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## Final Technical Report

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MILLIMETER WAVE TRANSMISSION  
SYSTEMS AND RELATED DEVICES

Abstract

by

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A survey has been made of the "state of the art" in millimeter (20 GHz - 300 GHz) wave transmission systems and related devices. The survey includes summaries of analytical studies and theoretical results that have been obtained for various transmission line structures. This material has been supplemented by further analysis where appropriate.

The transmission line structures are evaluated in terms of electrical performance, ease of manufacture, usefulness for building other devices and compatibility with solid state devices.

Descriptions of waveguide transmission lines which have commonly been used in the microwave frequency range are provided along with special attention given to the problems that these guides face when their use is extended into the millimeter wave range. Also, guides which have been introduced specifically to satisfy the requirements of millimeter wave transmission are discussed in detail.

## ACKNOWLEDGEMENTS

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## CHAPTER 1

### INTRODUCTION

The millimeter wave spectrum (from 20GHz-300GHz) has not been exploited for use as have the lower microwave frequencies for a number of reasons.

Technological barriers have held back development for a number of years but recent advances have eliminated major obstacles such as a lack of coherent sources and amplifiers.

Tube sources such as the gyrotron and solid state devices such as the IMPATT and GUNN diode oscillators can be made to produce substantial millimeter wave power. As these technologies develop the cost effectiveness should increase, encouraging further development of devices and systems.

The application of gallium arsenide FET's in solid state amplifiers has extended higher and higher in frequency and may be practical up to 60 GHz [1]. FET's based on other semiconductors, such as indium phosphide, will have potential application through the higher millimeter frequencies.

Another disadvantage to the development of millimeter wave frequencies is the effects of atmospheric absorption on propagation. Atmospheric "windows" in the frequency spectrum occur between peaks of molecular absorption "lines" and allow for maximum atmospheric penetration. These occur at 94, 140 and 230 GHz. Attenuation

due to rain is large because the size of a rain droplet is on the order of one wavelength in the millimeter wave region.

The cost and manufacturing difficulties which occur at higher frequencies due to the decrease in size of many components have been partially overcome by improved techniques in planar construction, such as ion-beam lithography and by the introduction of new guiding structures, based on optical wave techniques, whose operating properties are not critically dependent on small size and strict tolerances.

Perhaps the factor which has most discouraged the use of the millimeter wave spectrum is the lack of a significant large-scale system application [2]. Such an application, radar, existed to boost the development of the microwave frequencies following World War II. Although no single application is apparent for the millimeter wave frequencies, a number of potential projects exist that will open up that spectrum to the type of rapid development that occurs when methods and techniques give rise to even further applications.

The applications for the millimeter spectrum are diverse. Some applications, such as communication, are an extension of those for the microwave frequencies while other, such as imaging utilize the specific properties of these waves.

Millimeter wave communication systems, despite the problems in propagation through the atmosphere, are being developed for purposes which avoid or employ those effects.

Space-to-space communications occur above the atmosphere and due to the small millimeter wavelengths, satellites can employ small directional antennas to communicate with each other and thus increase the range of earth-space-earth communication links. Additionally, the natural cryogenic environment in space would allow for very low noise figures at millimeter wavelengths.

Since high gain antennas are available at a reduced size and due to the effects of the atmosphere, land-based, "secure", communication links are another application. Highly directed signals which do not carry for longer distances than necessary make for a low probability of intercept. These systems are attractive for military use.

Millimeter wave imaging systems are being examined for military use in target acquisition and fire control [3]. The low attenuation of millimeter waves due to battlefield smoke and ground clutter is cited as an advantage in this application.

Other areas of millimeter wave frequency application are in the studies of materials, in devices for detecting and treating tumors and in astronomy and remote sensing from space [4].

The combination of the improved technologies available and the increase in the number of potential applications will move to reduce the disparity between millimeter wave frequency usage and the usage of the lower microwave frequencies. In this paper the state of the art in millimeter wave transmission systems and

devices will be examined for electrical performance, ease of manufacture and cost effectiveness.

Chapter 2 describes closed metallic guides, which have been used extensively at microwave frequencies, with an explanation of their limitations at millimeter wave frequencies. Modifications of standard guides which increase their effectiveness in this range, such as finline and oversized guide, are also described.

Planar guides, which were introduced to decrease the size, weight and cost of microwave circuits are examined for potential use at millimeter frequencies in Chapter 3. Traditional microstrip and stripline are discussed as well as their variations.

Chapter 4 describes a new family of guides which have many advantages at millimeter-wave frequencies. Dielectric guides are not subject to extreme miniaturization and ohmic losses as are the previously described guides. Recent theoretical techniques developed for analysis of these guides are given.

Finally Chapter 5 briefly describes the components and devices formed from these guides which are useful for millimeter-wave circuits.

## CHAPTER 2

### CLOSED METALLIC WAVEGUIDES

#### 2.1 Introduction

Closed metallic guides have been used for years as a medium for microwave frequency transmission because they overcome many of the mechanical and electrical disadvantages of the structures used at lower frequencies, such as coaxial lines and open two-conductor lines. Coaxial transmission lines and open two-conductor lines require dielectric supports to keep the two conductors uniformly spaced. The use of two conductors and insulating support material results in considerably higher attenuation than that of hollow tubes or guides. Another significant shortcoming of open two-conductor transmission lines is that because of no natural shielding it is necessary to keep the transmission line well away from surrounding objects in order to avoid unwanted reflections and stray radiation. Also all bends must have a large radius of curvature in order to reduce radiation that occurs from such bends. These are the main reasons why such lines are not generally used at millimeter wavelengths for general transmission over long distances. Short lengths of coaxial transmission lines may be used at the longer millimeter wavelengths to couple circuits together. Extending the use of closed guides into the millimeter wave frequencies is natural since many of the advantages are

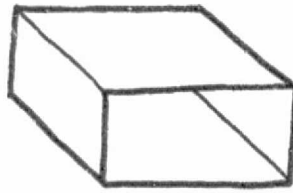
retained into this range. Design and construction of millimeter guides and circuits can be treated as a matter of scaling down microwave guides. Guides with appropriately reduced cross sectional openings are available for use at frequencies as high as 325 GHz. At 325 GHz the guide width is only 0.8636 mm.

As the frequency increases, new problems are encountered, notably conduction losses increase due to the skin effect. Tolerances, which are held to a percentage of guide dimensions, can be met only at considerable increases in cost of manufacturing.

In the following sections, closed metal guides are discussed in detail with consideration to the problems encountered in the use at millimeterwave frequencies and some of the steps that have been taken to overcome them. The cross section of the most common closed metallic guides are shown in Fig. 2.1.

## 2.2 Losses

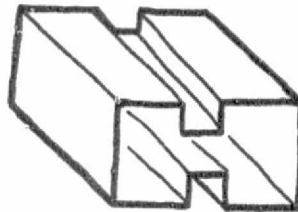
The predominant mechanism leading to attenuation of energy propagating in closed metallic guides is ohmic loss in the conducting walls. This loss depends on the size and shape of the guide's cross section, the field distribution of the propagating mode and the frequency of operation. This last dependence is of most concern at millimeter-wave frequencies and is described by the skin effect.



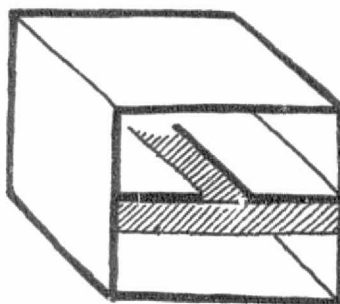
a) Rectangular



b) Circular



c) Ridged



d) Finline

Figure 2.1 Cross Sections of Various Closed Metallic Guides

### The Skin Effect

Fields at the surface of a good conductor produce in that conductor a current with a density that decreases exponentially with increased distance from the surface. The rate at which the current density falls off increases with the square root of the frequency of the exciting fields so at very high frequencies most of the current flows in a very thin layer near the conductor's surface. This is called the skin effect and is described quantitatively by the skin depth,  $\delta_s$ , which is the distance at which the current density falls to  $1/e$  of its value at the surface. The skin depth depends on the physical parameters of the conductor and the frequency:

$$\delta_s = (\pi\mu\sigma f)^{-\frac{1}{2}} \quad (2.1)$$

where  $\mu$  is the permeability,  $\sigma$  is the conductivity, and  $f$  is the frequency.

The skin depth is defined for solid conducting planes but conducting surfaces with a thickness and radius of curvature of several skin depths are accurately described by this quantity. The skin depth of several common conductors as a function of frequency is given in Fig. 2.2. At 100 GHz the skin depth is of order  $5 \times 10^{-5}$  cm., i.e. less than one micron.

The result of the skin effect is to increase the resistance of the conducting surface by effectively reducing the area



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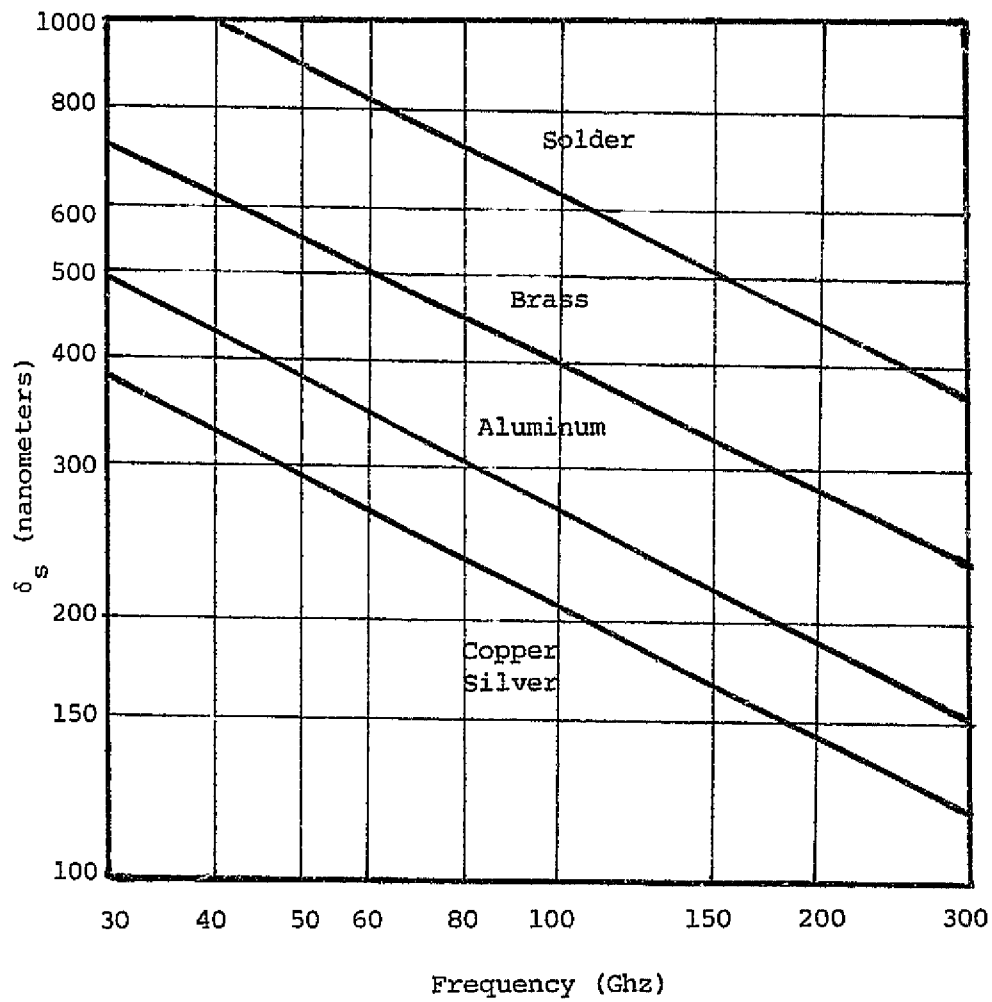


Figure 2.2 Skin Depth vs. Frequency for Several Plane  
Conductors

through which the currents may flow and thus ohmic losses are increased. The surface resistance is equal to  $(\sigma\delta_s)^{-1}$  and so increases with the square root of the frequency.

Several methods to reduce the skin effect have been proposed. These are described below.

Guides can be constructed of composite materials made up of thin conducting sheets placed parallel to one another and insulated by thin dielectric layers as shown in Fig. 2.3.

If the laminations are arranged parallel to the surface of the guide as in Fig. 2.3a and the conducting layers are thin compared to the skin depth, the fields may tunnel deeper into the structure and increase the area available for conduction currents [5].

Alternatively, the layered structure can be used to construct a guide with the laminations perpendicular to the direction of power flow, as in Fig. 2.3b. Currents will exist in the dielectric layers at a depth much greater than the skin depth, and if the conducting layers are very thin, the current density will be uniform along the guide length [6].

Using these methods, it is possible to reduce the negative results of the skin effect in solid conductors, but this depends on the ability to construct the laminate materials to a high degree of accuracy and uniformity. This critical dependence on the dimensions reduces the bandwidth and increases the cost of these guides making them impractical for most applications.

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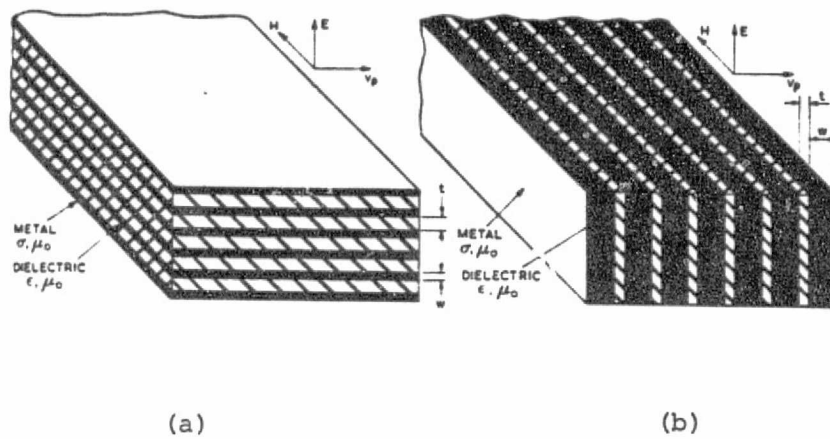


Figure 2.3 Two Laminated Structures Proposed to Reduce the Skin Effect, (a) parallel with and (b) perpendicular to the direction of propagation.

(Reproduced from Harvey, A.F. Microwave Engineering. London: Academic Press, 1963).

### Surface Effects

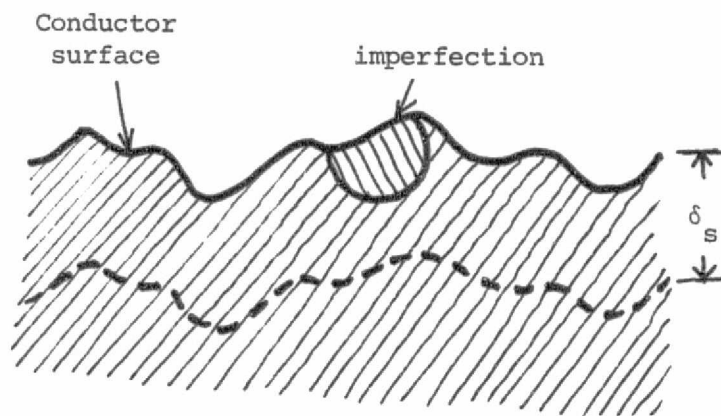
While the skin effect increases ohmic losses at high frequencies by reducing the area through which conduction currents can flow, it has a secondary effect of increasing the sensitivity of the ohmic losses to the nature of the surface itself. This can be seen in Fig. 2.4. As the skin depth decreases to the order of magnitude of the surface imperfections, these imperfections make up a large part of the conduction current path. This leads to a value of surface resistance,  $R_s$ , at millimeter frequencies that differs from the customary theoretical value determined with consideration only to the primary effects of reduced skin depth, as in the previous section.

This result has been studied experimentally [7] and a relationship between theoretical and experimental values of  $R_s$  has been determined which is useful in finding the attenuation of waveguides and the  $Q$  factors of resonators and circuits.

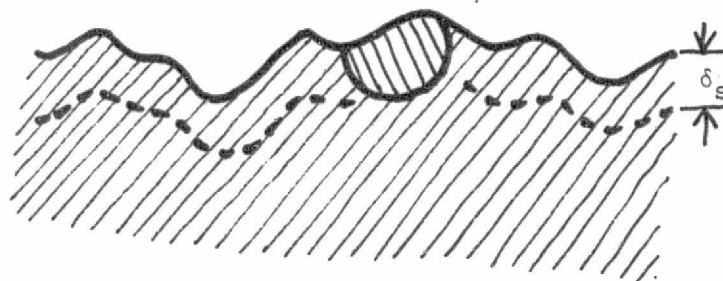
The results of these experiments indicate that an increase in surface resistance from theoretical values is the result of three different surface effects.

### Surface Roughness

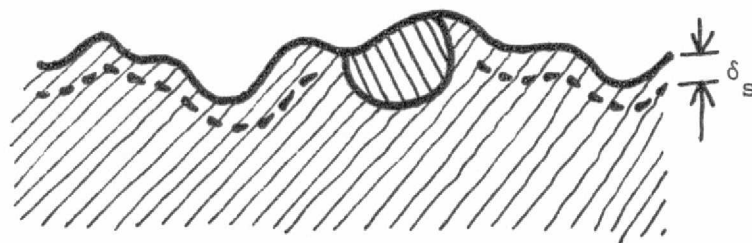
The effects which potentially leads to the largest deviation from theory is surface roughness as seen in Fig. 2.5. Surface roughness effectively increases the conduction path length from a length,  $l$ , which is the actual length of a perfectly flat surface,



(a)



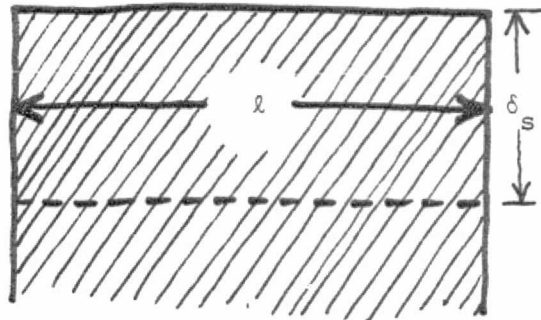
(b)



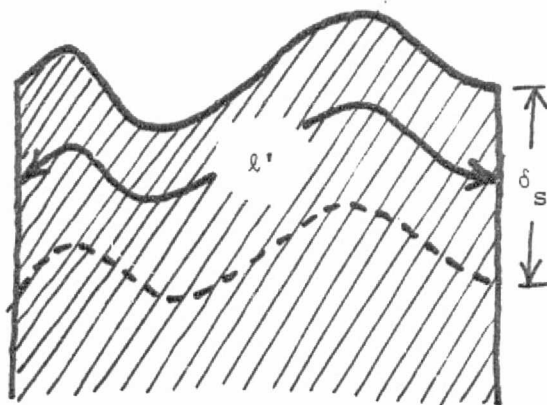
(c)

Figure 2.4 The Effect of Skin Depth on the Surface Current Path

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(a)



(b)

Figure 2.5 The Increase in Current Path Length with  
Increasing Surface Roughness

to a length  $\ell'$ , which is the average distance traveled by currents along the rough surface of the same actual length. Surface roughness is, of course, a two dimensional property and methods of manufacture and finishing of metal guides often create a surface roughness which varies with direction along the surface. Drawn sections of guides often have very little roughness in the longitudinal direction, while machining operations such as turning or boring create a roughness which is greatest perpendicular to the lay produced by the operation. This implies that, when possible, machining operations required for millimeter waveguide components should be carried out so that the lay is parallel to the direction of r.f. current flow. Increases in surface resistance over theoretical values are shown in Fig. 2.6 [7] and may be as large as 30%. Surface roughness is usually expressed as a root mean squared (r.m.s.) or center line average (c.l.a.) height. These values are practically equal. Surface roughness has an effect when it is larger than the skin depth. Since the effect is an increase in ohmic losses it is desirable to keep the roughness less than half of the skin depth. For most conductors at millimeter wave frequencies a value between 0.05 and 0.2  $\mu\text{m}$  would usually be specified. Surface roughness can be measured mechanically with a stylus-type profilometer fitted with a stylus of a suitably small radius, or by the electrical output of a pickup travelling across the surface. A cross section of the surface may be photographed

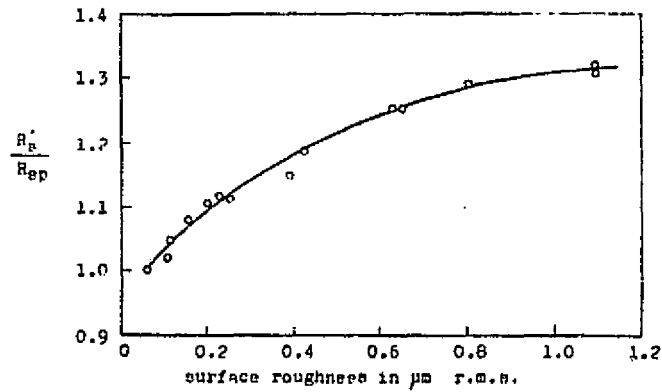


Figure 2.6 Fractional Increase in Surface Resistance over  
Calculated Values vs. Surface Roughness

(Reproduced from Tischer, Frederick J. "Excess Conduction Losses at Millimeter Wavelengths." IEEE Trans. MTT, Vol. MTT-24, No. 11 (Nov. 1976), pp. 853-58).



and enlarged and the surface roughness measured directly from the photograph. Alternatively the interference pattern of monochromatic light reflected from the surface and from an optically flat plate can be calibrated to determine roughness.

Surface roughness can be reduced by several methods. Etching and annealing improve metal surfaces but dimensional changes must be accounted for. Electropolishing is based on the preferred dissolution of surface irregularities in an electrolyte under the influence of a current. This process can provide good surfaces for use at millimeter wavelengths. Buffing and polishing produce smooth surfaces but can result in a low conduction layer above the metal which can increase losses.

Overall, care must be taken to assemble waveguide circuits and components. Careful handling in a clean environment will preserve the advantages gained by creating a smooth surface.

#### Work Hardening Effects

When the crystal structure of a metal is deformed below the recrystallization temperature of that metal, the result is called work hardening. Such deformation can occur at the surface of guide components while they are being formed, machined or mechanically polished. As a result, the conductivity within the skin depth is decreased and the surface resistance increases. It has been found experimentally that work hardening for mechanically highly polished surfaces increases  $R_s$  by a factor of 1.18 from its calculated value [7].

### Anomalous Skin Effect

Experiments carried out at millimeter frequencies [8] give a value of surface resistance for metals, with which extreme care has been taken to provide the lowest possible roughness, that are considerably higher than the value predicted by theory. These measurements have lead to a theoretical explanation for what has been called the anomalous skin effect in conductors at millimeter and submillimeter wavelengths [9]. The anomalous effect is explained by the dependence of the conductivity,  $\sigma$ , on the frequency and field wave-number due to charge density fluctuations at the surface. These fluctuations are due to the fields normal to the surface. The theory agrees with the experimental results and is applicable to normal metals used for millimeter waveguides. For copper, the anomalous skin effect produces an increase of 13.5% in the surface resistance,  $R_s$ , at 35 GHz and an increase of 20% at 70 GHz.

### 2.3 Rectangular Waveguides

The principle of operation for all waveguides is the propagation of electromagnetic waves along their length. For closed metallic guides this propagation can occur for fields which have either an electric or a magnetic field component in the longitudinal direction. These are called transverse magnetic (TM) or transverse electric (TE) modes. An infinite number of discrete field configurations, or modes, are possible for each type. At any

given frequency only a finite number of modes can propagate. For rectangular waveguides, modes are identified by integer subscripts,  $m$  and  $n$ , for each type and referred to as  $TE_{mn}$  or  $TM_{mn}$  modes. In accordance with Maxwell's equations and the boundary conditions at the metal surface, the transverse fields vary sinusoidally within the guide in the directions of its height and width. The subscripts  $m$  and  $n$  are the number of half-wavelengths in each of these directions respectively. The fields are periodic in the longitudinal direction with a period,  $\lambda_g$ , called the guide wavelength. There exists a frequency,  $f_c$ , below which propagation cannot occur. This frequency depends on the guide dimensions. The lowest propagating mode is a TE mode.

Choosing guide dimensions for a particular frequency use depends on many factors including bandwidth, attenuation and power-handling capability. Bandwidth is limited by the low frequency cutoff, and the sensitivity of guide impedance and attenuation near that cutoff, and multimode propagation at higher frequencies. The attenuation varies with frequency, but the variation depends on the mode of propagation. The skin effect, as described earlier, increases the surface resistance of the guide in proportion to the square root of the frequency, independent of the guide dimensions.

The attenuation for rectangular guides can be calculated from the field expressions using the perturbation method [10]. The

attenuation for the dominant  $TE_{10}$  mode is

$$\alpha = \frac{R_s}{ab\beta_o k_o Z_o} (2bk_c^2 + ak_o^2) \frac{\text{nepers}}{m} \quad (2.2)$$

where  $R_s$  is the surface resistance ( $R_s = (\pi f \mu / \sigma)^{\frac{1}{2}}$ ),  $k_c$  is the cutoff wavenumber ( $k_c = \pi/a$ ) for the  $TE_{10}$  mode,  $\beta$  is the propagation factor of the  $TE_{10}$  mode,  $Z_o$  is the impedance of free space and  $a$  and  $b$  are the wide and narrow dimensions of the guide, respectively.

The attenuation for rectangular guides with various aspect ratios, operating in the  $TE_{10}$  mode, is given as a function of frequency in Fig. 2.7 [11]. At 90 GHz attenuation is about 3 dB/m.

The maximum power that can be transferred in the guide increases with the cross-sectional area of the guide and the breakdown field of the material filling the guide. Thus, the miniature guides used at millimeter wave frequencies have reduced power handling capability compared with the large guides used at lower frequencies. The use of high field breakdown gases, such as Freon and sulfur hexafluoride, introduced in sealed guide sections, increases the maximum power that can be used in those sections. But for millimeter wave guides the advantages so gained must be weighed against possible formation of non- or low-conductive layers and corrosion at the metal surface due to the gas. At these frequencies the skin effect amplifies the negative effect these imperfections have on attenuation.

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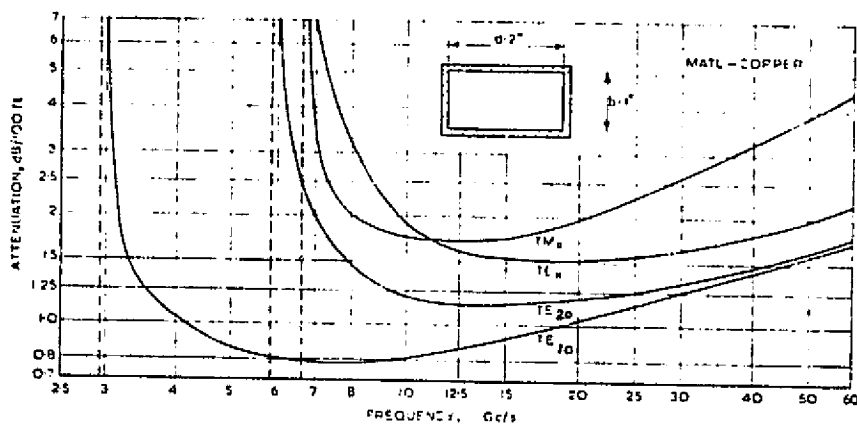


Figure 2.7 Rectangular Guide Attenuation vs. Frequency

(Reproduced from Ramo, Simon; Whinnery, John R.; and Van Duzer, Theodore. Fields and Waves in Communication Electronics. New York: John Wiley & Sons, Inc. 1965).

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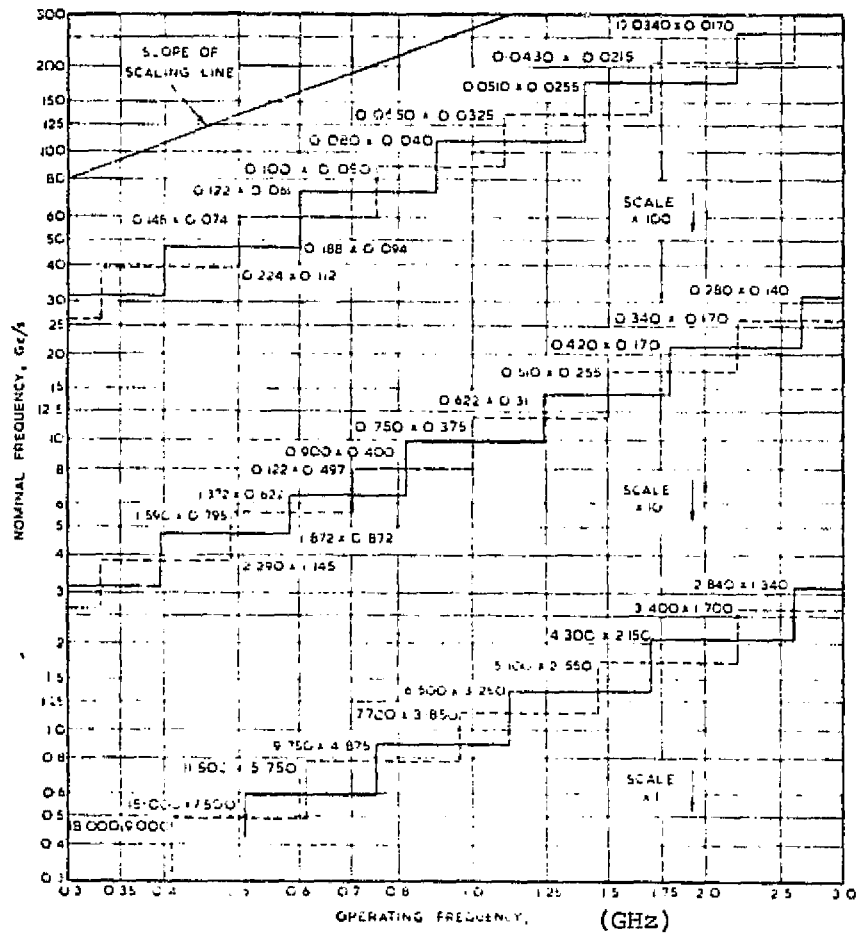


Figure 2.8 Rectangular Guide Ranges

(Reproduced from Harvey, A. Microwave Engineering.  
London: Academic Press, 1963).

With respect to these considerations a standard set of rectangular waveguides exists which covers the various frequencies. The ranges covered by each guide are given in Fig. 2.8 [12].

These standards are based on a  $TE_{10}$  mode of propagation with a width-to-height ratio of 2.0. For these guides the operating frequency is usually  $1.5 f_c$  where  $f_c = \frac{c}{2a}$  for these modes ( $c$  is the speed of light). The recommended frequency limits of operation for these guides are  $\pm 20\%$  of this nominal operating frequency due to the factors discussed above. Cutoffs for some guide modes, normalized to the dominant mode cutoffs, are shown in Fig. 2.9. Technical data for the standard guides, which are designated by numbered types are given in Fig. 2.10 [12]. Guides WG-19 through WG-32 cover the millimeter wave spectrum.

The power ratings for air-filled guides with a breakdown field of 30 kV/cm and a safety factor of four, and the attenuation for guides made of pure copper are calculated from theory for a frequency of  $1.5 f_c$ .

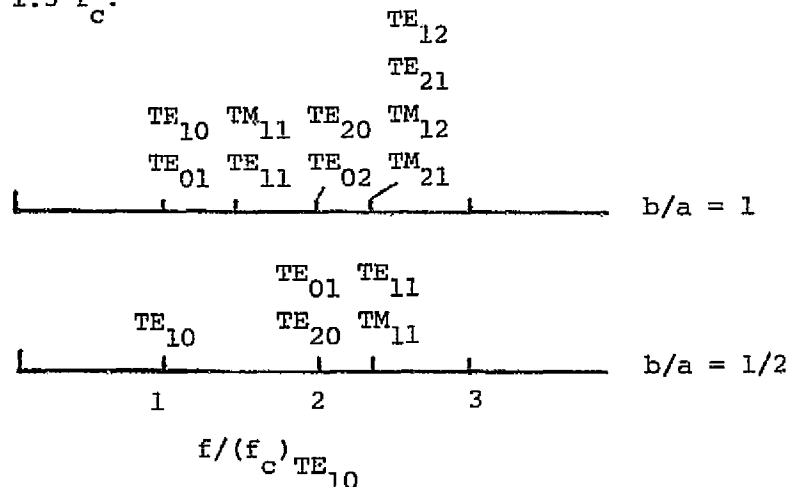


Figure 2.9 Relative Frequency Cutoffs in Standard Rectangular Guide

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*Technical Data for Standard Rectangular Guides*

British R.C.S.C. type	American E.I.A. type	Inside dimensions, in		TE <sub>01</sub> cut-off, Gc/s	Operating range, Gc/s	Attenua- tion, dB/100 ft	Power rating, kW
		Width	Depth				
WG 1	WR 1800	18.000	9.000	0.327	0.41-0.61	0.0334	115,000
WG 2	WR 1500	15.000	7.500	0.393	0.51-0.75	0.0438	85,000
WG 3	WR 1150	11.500	5.750	0.517	0.61-0.96	0.0614	45,000
WG 4	WR 975	9.750	4.875	0.605	0.75-1.12	0.0807	33,000
WG 5	WR 770	7.700	3.850	0.765	0.96-1.45	0.120	21,500
WG 6	WR 650	6.500	3.250	0.908	1.12-1.70	0.154	14,800
WG 7	WR 510	5.100	2.550	1.154	1.45-2.20	0.222	9,300
WG 8	WR 430	4.300	2.150	1.375	1.70-2.60	0.286	6,500
WG 9 A	WR 340	3.400	1.700	1.737	2.20-3.30	0.450	4,200
WG 10	WR 284	2.840	1.340	2.080	2.60-3.95	0.555	2,800
WG 11 A	WR 229	2.290	1.145	2.579	3.30-4.90	0.750	2,000
WG 12	WR 187	1.872	0.872	3.155	3.95-5.85	1.047	1,200
WG 13	WR 159	1.590	0.795	3.710	4.90-7.05	1.259	900
WG 14	WR 137	1.372	0.622	4.285	5.85-8.20	1.700	620
WG 15	WR 112	1.122	0.497	5.260	7.05-10.0	2.338	420
WG 16	WR 90	0.900	0.400	6.56	8.20-12.4	3.24	250
WG 17	WR 75	0.750	0.375	7.87	10.0-15.0	3.92	205
WG 18	WR 62	0.622	0.311	9.49	12.4-18.0	5.21	145
WG 19	WR 51	0.510	0.255	11.57	15.0-22.0	7.00	95
WG 20	WR 42	0.420	0.170	14.08	18.0-26.5	10.9	52
WG 21	WR 34	0.340	0.170	17.37	22.0-33.0	12.8	43
WG 22	WR 28	0.280	0.140	21.10	26.5-40.0	17.3	28
WG 23	WR 22	0.224	0.112	26.35	33.0-50.0	24.0	18
WG 24	WR 19	0.188	0.094	31.4	40.0-60.0	31.3	14
WG 25	WR 15	0.148	0.074	39.9	50.0-75.0	44.7	8
WG 26	WR 12	0.122	0.061	48.4	60.0-90.0	59.7	5
WG 27	WR 10	0.100	0.050	59.0	75.0-110	80.7	3.8
WG 28		0.0800	0.0400	73.8	90.0-140	113	2.3
WG 29		0.0650	0.0325	90.9	110-170	154	1.5
WG 30		0.0510	0.0255	115.8	140-220	225	0.90
WG 31		0.0430	0.0215	137.5	170-280	286	0.65
WG 32		0.0340	0.0170	173.3	220-325	405	0.42

Figure 2.10 Technical Data for Standard Rectangular  
Guides.

(Reproduced from Harvey, A.F. Microwave Engineering.  
London: Academic Press, 1963.)



## 2.4 Circular Waveguides

Circular waveguides have the same modal properties that rectangular guides have, that is there exist an infinite number of discrete modes of the TE and TM types. Two integer subscripts identify the modes,  $TE_{mn}$  or  $TM_{mn}$ , where  $m$  and  $n$  are the number of half cycles of the transverse field components in the circumferential and radial directions, respectively. The fields are periodic in the longitudinal direction with a period,  $\lambda_g$ , called the guide wavelength, and as with the rectangular guide each mode has a low frequency cutoff that depends on the guide dimensions. The lowest and dominant propagating mode is the  $TE_{11}$  mode.

Circular guides that operate with the dominant mode have a disadvantage in that, due to the circular symmetry of the guide, there exists no preferred direction for the plane of electric field polarization so that discontinuities result in rotation of the fields. This effects the operation of components such as mode transducers which rely on a specific direction of polarization.

Modes with circular symmetry,  $TE_{0n}$  and  $TM_{0n}$ , overcome this disadvantage. Additionally,  $TE_{0n}$  modes have the property that attenuation due to ohmic losses decreases with increasing frequency. This makes circular guides operating with a  $TE_{0n}$  mode attractive for use at millimeter wave frequencies, especially for trunk communication systems where low attenuation of signals travelling over a long distance is required.

In order to insure single mode propagation of the  $TE_{on}$  mode, filters must be used to attenuate unwanted modes excited at bends and discontinuities in the line. Since the  $TE_{on}$  modes are the only modes which have no axial component of conduction currents flowing in the guide walls, circular guides with walls formed by a tightly wound helical coil with a low pitch angle cut the axial current path and non- $TE_{on}$  modes are thus rapidly attenuated.

The expression for attenuation of the low-loss  $TE_{on}$  modes is

$$\alpha = \frac{R_s}{aZ_o} \frac{f_c^2}{f(f^2 - f_c^2)} \frac{\text{nepers}}{m} \quad (2.3)$$

where  $f_c$  is the low cutoff frequency of the  $TE_{on}$  mode and  $a$  is the radius of the guide. The attenuation for circular guides operating in the  $TE_{11}$  and  $TE_{01}$  modes is given in Fig. 2.11 [11].

At 10 GHz the attenuation of these modes is the same, about 0.015 dB/m for 2" diameter pipe, while at 40 GHz the attenuation increases to 0.4 dB/m for the  $TE_{11}$  mode and decreases to 0.0015 dB/m for the  $TE_{01}$  mode, for the same size pipe.

Standard sizes for circular guides are chosen to match frequency bands with standard rectangular guides and are adjusted to agree with commercially available machine tools. Technical data for these standard guides is given in Fig. 2.12 [12].

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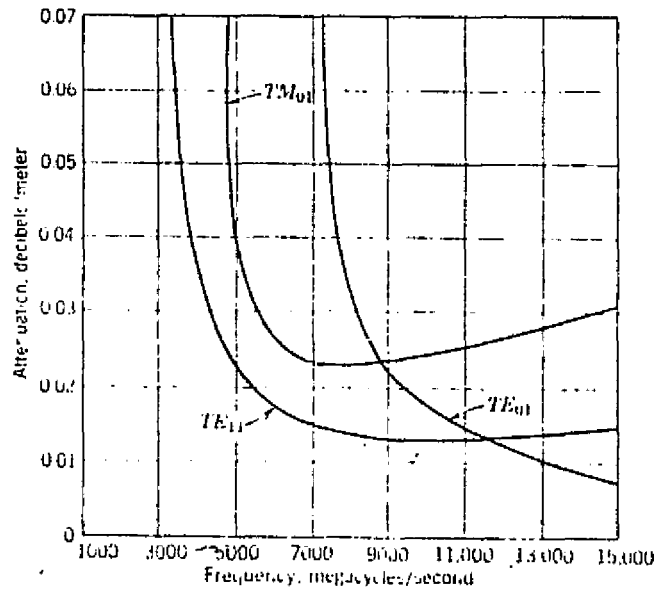


Figure 2.11 Circular Wave Guide Attenuation  
vs. Frequency

(Reproduced from Ramo, Simon; Whinnery, John R.; and Van Duzer, Theodore. Fields and Waves in Communication Electronics. New York: John Wiley & Sons, 1965).

(All dimensions are in inches)

American E.I.A. type	Inside section			Nominal outside diameter	Wall thickness		Frequency range $TE_{11}$ mode, Gc/s	Frequency range $TE_{01}$ mode, Gc/s
	Diameter	Tolerance + or -	Roundness tolerance		Mean	Tolerance + or -		
WC 2551	25.508	.025	.025				0.312-0.427	0.683-0.940
WC 2179	21.791	.020	.022				0.365-0.500	0.799-1.10
WC 1862	18.616	.020	.019				0.427-0.586	0.936-1.29
WC 1590	15.903	.015	.016				0.500-0.686	1.10-1.51
WC 1359	13.585	.015	.014				0.586-0.803	1.28-1.77
WC 1161	11.606	.010	.012				0.686-0.939	1.50-2.07
WC 892	9.915	.010	.010				0.803-1.10	1.76-2.42
WC 847	8.470	.008	.008	8.630	.080	.008	0.939-1.29	2.06-2.83
WC 724	7.235	.007	.007	7.395	.080	.008	1.10-1.51	2.41-3.31
WC 618	6.181	.006	.006	6.241	.080	.008	1.29-1.76	2.82-3.88
WC 528	5.280	.005	.005	5.440	.080	.008	1.51-2.07	3.30-4.54
WC 451	4.511	.005	.005	4.671	.080	.008	1.76-2.42	3.86-5.32
WC 385	3.853	.004	.005	4.013	.080	.008	2.07-2.83	4.52-6.22
WC 329	3.292	.003	.003	3.452	.080	.008	2.42-3.31	5.29-7.28
WC 281	2.812	.003	.003	2.942	.065	.006	2.83-3.88	6.19-8.53
WC 240	2.403	.0025	.002	2.533	.065	.005	3.31-4.54	7.25-9.98
WC 205	2.047	.002	.002	2.177	.065	.004	3.89-5.33	8.51-11.7
WC 175	1.750	.0015	.0015	1.880	.065	.003	4.54-6.23	9.95-13.7
WC 150	1.500	.0015	.0015	1.630	.065	.003	5.30-7.27	11.6-16.0
WC 128	1.281	.0013	.0013	1.411	.065	.0025	6.21-8.51	13.6-18.7
WC 109	1.094	.001	.0011	1.194	.050	.002	7.27-9.97	15.9-21.9
WC 94	0.938	.0009	.0009	1.038	.050	.002	8.49-11.6	18.6-25.6
WC 80	0.797	.0008	.0008	.897	.050	.0015	9.97-13.7	21.9-30.1
WC 69	0.688	.0007	.0007	.788	.050	.0015	11.6-15.9	25.3-34.9
WC 59	0.594	.0006	.0006	.674	.040	.001	13.4-18.4	29.3-40.4
WC 50	0.500	.0005	.0005	.580	.040	.001	15.9-21.8	34.8-48.0
WC 44	0.438	.00045	.0004	.518	.040	.001	18.2-24.9	39.8-54.8
WC 38	0.375	.00038	.0004	.435	.030	.001	21.2-29.1	46.4-63.9
WC 33	0.328	.00033	.0003	.388	.030	.001	24.2-33.2	53.1-73.1
WC 28	0.281	.00028	.0001	.341	.030	.001	28.3-38.8	61.9-85.2
WC 25	0.250	.00025	.0001	.290	.020	.001	31.8-43.6	69.7-95.9
WC 22	0.219	.00025	.0001	.259	.020	.001	36.4-49.8	79.6-110
WC 19	0.188	.00025	.00007	.228	.020	.001	42.4-58.1	92.9-128
WC 17	0.172	.00025	.00007	.212	.020	.001	46.3-63.5	101-139
WC 14	0.141	.00025	.00005	.181	.020	.001	56.6-77.5	124-171
WC 13	0.125	.00025	.00005	.155	.015	.001	63.5-87.2	139-192
WC 11	0.109	.00025	.00005	.139	.015	.001	72.7-99.7	159-219
WC 9	0.094	.00025	.00005	.124	.015	.001	84.8-116	186-256

Figure 2.12 Technical Data for Standard Circular Guides

(Reproduced from Harvey, A.F. Microwave Engineering.  
London: Academic Press, 1963).

## 2.5 Ridged Waveguides

Ridged waveguides are useful at microwave frequencies to reduce the size and increase the bandwidth of conventional rectangular guides. At millimeter wave frequencies these advantages become disadvantages due to the already too small dimensions of the guide. The ridged shape increases the construction difficulty over the relatively simple rectangular shape and the small spacing between the ridges of the guide further reduces the power handling capability.

For these reasons standard ridged waveguides cover only the frequency range of 1-40 GHz. At 40 GHz, a standard double ridged guide with an operating bandwidth of 2.4 to 1 has overall inside dimensions of 0.288" x 0.134" and a ridge spacing of 0.057" [12].

The attenuation for the same guide is 0.72 dB/m, which is the same as that of a rectangular guide operating at 40 GHz.

## 2.6 Finline

Finline transmission lines overcome some of the disadvantages of the previously described closed metallic guides while retaining the advantage of being a closed guide.

Typical finline structures can be viewed as shielded slotline, dielectric-loaded ridged waveguide, or dielectric-loaded rectangular waveguide with fins. Consistent with these views, finline, like ridged waveguide, has wide bandwidth capabilities and can be designed to give single-mode octave bandwidths.

In most finline structures metal fins are printed on a dielectric substrate which bridges the broad walls of a rectangular waveguide. This construction technique utilizes existing low-cost batch photoetching techniques which increases the ability of finline to be useful in forming passive hybrid integrated circuits.

If the dielectric substrate is allowed to pass entirely through the walls of the rectangular guide and an additional dielectric spacer is placed between one of the fins and the guide, complete dc isolation of one fin is provided for the biasing of solid-state devices mounted between the two fins. This arrangement is shown in Fig. 2.13. The thickness of the guide walls at the gaps in these walls are one quarter of a wavelength so that they still provide good rf shielding. The ease with which solid-state devices can be mounted is a distinct advantage over the closed metallic guides described previously.

Finline is useful at millimeter wavelengths since the advantages of planar construction, as described above, are realized without the miniaturization required for standard, unshielded planar guides like slotline and microstrip. These guides require substantially reduced substrate thicknesses or high dielectric constants to reduce higher mode propagation and radiation as frequency is increased to the millimeter wave range. This leads to reduced Q factors, critical tolerances and mechanically weak components. Since finelines are closed structures, radiation is

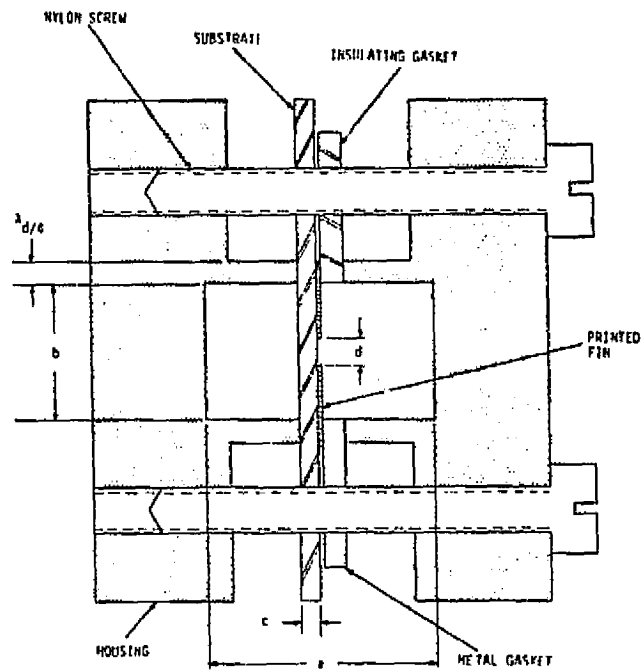


Figure 2.13 Finline Structure

not a problem and extreme miniaturization and high dielectric constant materials are not required.

The rectangular guide used for shielding in finlines are the same size as standard rectangular guides for operation at a given frequency in the millimeter wave frequency band. Thus, overall finline dimensions are subject to the same problems in miniaturization as standard rectangular guides although this is an advantage when transitions between finline and standard rectangular guide are required.

Attenuation in finline is smaller than that of microstrip at millimeter wave frequencies since there is no radiation loss and low-loss dielectrics can be used. This has been demonstrated experimentally [13].

The dispersion of finlines is nearly that of a dielectric loaded rectangular guide for large fin separations and can be calculated using the transverse resonance technique. For small fin separations ( $w/D \ll 2$ ) the finline behaves like a slotline and the guide walls have little effect on the propagation characteristics of the guide. For finlines which utilize low-dielectric-constant substrates there is an approximation for the dispersion relation which is

$$\frac{\lambda'}{\lambda} \approx \frac{1}{\sqrt{k_e - (\lambda/\lambda_c)^2}} \quad (2.4)$$



where  $\lambda$  is the free space wavelength,  $\lambda_c$  is the low frequency cutoff wavelength,  $\lambda'$  is the guide wavelength, and  $k_e$  is the experimentally determined effective dielectric constant of the structure and is invariant with frequency.

The dispersion characteristics given as  $\lambda'/\lambda$  as a function of frequency are given for some typical finline structures in Fig. 2.14 [14].

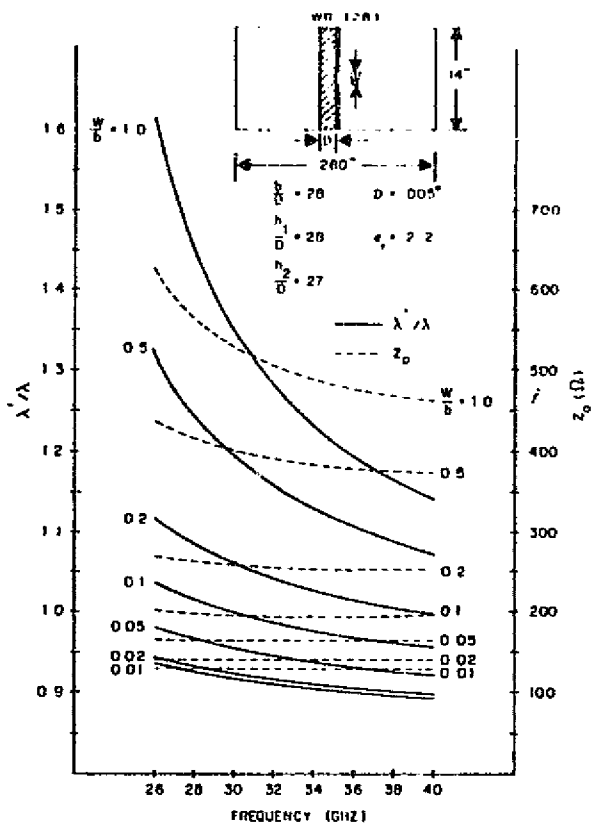
## 2.7 Oversized Waveguides

Standard rectangular waveguides are reduced in size as frequency increases to insure single mode propagation. Multimode propagation leads to signal distortion since the phase velocity of the various propagating modes are different.

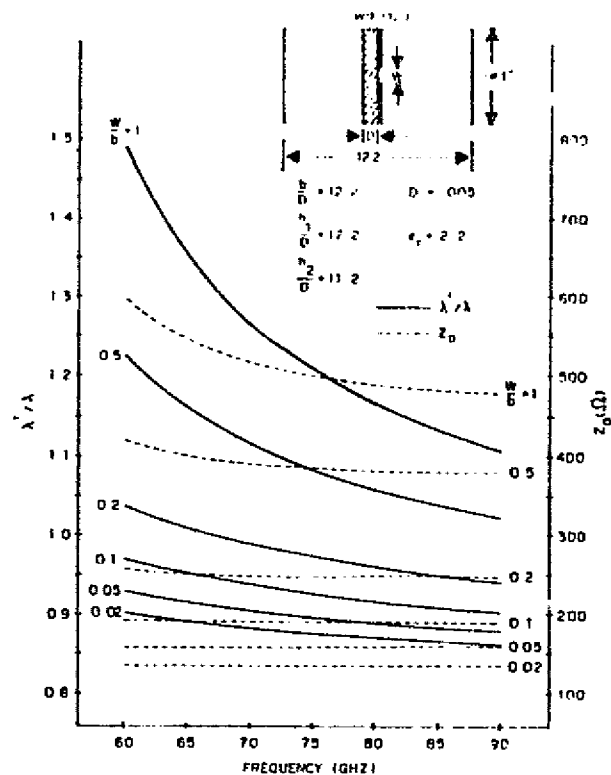
This reduction in size has many disadvantages. Miniaturization leads to critical tolerances and reduced surface area of the metal walls causes increased attenuation due to ohmic losses in the metal, decreased power handling capability and decreased heat dissipation at high powers.

At millimeter wave frequencies these disadvantages become so pronounced that oversized, overmoded waveguides can be used to some advantage.

Most of these guides are rectangular guides operating with the electric field vector of the dominant  $TE_{10}$  mode parallel to the broad wall. Oversized guide sections are usually coupled through



(a)



(b)

Figure 2.14 Finline Guide Parameters. Dispersion in terms of wavelength ratio and characteristic impedance for two finline guides. (a and b).

(Reproduced from Knorr, Jeffrey B., and Shayda, Paul M. "Millimeter-Wave Fin-Line Characteristics", IEEE Trans. MTT, Vol. MTT-28, No. 7 (July 1980, pp. 737-43.

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tapered transitions to standard-sized waveguides at the sending and receiving ends. This means that the unwanted, higher-order modes can exist only in the oversize section, since they are below cutoff in the standard guide. The transition sections can be designed to give a minimum of mode conversion from the desired mode to the higher order modes, but a finite amount of this conversion occurs. The unwanted modes are then "trapped" in the oversized sections and if the length of this section together with the transition section is an integral number of guide wavelengths of one of the trapped modes, this region will be resonant for that mode and transmitted power will be substantially reduced. This effect can be the most detrimental one in using oversized guides and can be minimized by careful design of transition tapers to minimize mode conversion at the sending and receiving ends of the guide and by eliminating discontinuities along the oversized guide section. These measures also reduce the distortion due to non-resonant mode conversion.

Analysis of the propagation in oversized guides is exactly the same as that used for standard rectangular guides and attenuation, impedance and dispersion characteristics can be determined in this way.

The reduction of attenuation in oversized waveguide compared to standard waveguide can be seen by example of calculated attenuations at various frequencies shown below:

TABLE 2.1  
Comparison of Standard and Oversized Guide Attenuation

f (GHz)	Standard Guide		Oversized Guide	
	Guide Dimensions (inches)	Theoretical Attenuation (dB/m)	Guide Dimensions (inches)	Theoretical Attenuation (dB/m)
35	0.280 × 0.140	0.52	6.0 × 6.0	0.007
70	0.148 × 0.074	1.33	0.90 × 0.40	0.155
140	0.0650 × 0.0325	5.0	0.90 × 0.40	0.227
300	0.0340 × 0.0170	12	0.90 × 0.40	0.32

Reductions in attenuation of the order of magnitude of that observed above have also been measured [15].

Traditionally, methods of transmission in which the aperture widths are very large compared to the free space wavelength are termed "quasi-optical". In oversized waveguides the transverse dimensions are on the order of one hundred wavelengths and the dominant  $TE_{10}$  mode of propagation begins to behave very much like a TEM wave over a significant portion of the guided wavefront. For this reason quasi-optical techniques for constructing components in oversized waveguides must be used. These components consist of planar structures placed in the guide so as to act uniformly across the wavefront such as dielectric prisms or metal gratings. Reactive elements, such as those used in standard waveguide which are formed from transmission line stubs and metallic vanes and posts, cannot be used in oversized guide since they would easily excite the unwanted, high order modes. For this reason, oversized guides are limited in use in microwave

circuit design and are more applicable to transmission of signal energy from point-to-point such as from antenna-to-receiver links in communication systems.

## CHAPTER 3

### PLANAR WAVEGUIDES

#### 3.1 Introduction

Planar waveguides present a low-cost, lightweight alternative to closed metallic waveguides at microwave frequencies. The simplicity of planar construction techniques provides the capability to mass produce complicated circuits economically. Considerable size reduction is an advantage in designing small packaged circuits and high performance distributed circuit elements such as directional couplers and filters.

As with closed metallic guides, the use of planar guides is extended into the millimeter frequency range by scaling guide dimensions inversely to frequency. Since these guides already represent a size reduction, the dimensions of the structures used in the millimeter ranges become critically small. Unlike closed metallic guides the field of propagating modes are not confined and discontinuities in the transmission line can lead to losses through the mechanisms of radiation and propagation of surface modes perpendicular to the direction of the line.

Ohmic losses occur in the metal sheets of these guides just as in the walls of the closed metallic guides. The thicknesses of the sheets commonly used are much greater than the skin depth.

The geometry of the conducting sheets in planar guides is such that dielectrics must be used to mechanically support them. The finite conductivity or loss tangent of the dielectric used causes an additional loss in these guides.

The introduction of dielectric layers alters the fields in guides that would normally support a TEM mode of propagation so that they include some axial component. Such guides are the strip-line, microstrip and coplanar guides which are said to support quasi-TEM modes of propagation. The cross section of these guides is shown in Fig. 3.1.

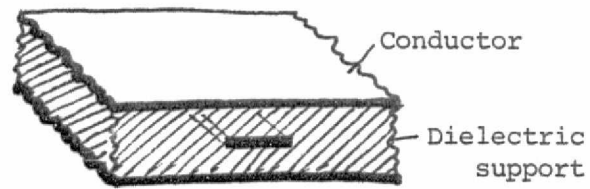
Other guides such as the slotline, shown in Fig. 3.1d, support a non-TEM mode which is comparable to that of finline guides.

The frequency separation between the propagating modes in these lines is such that they have a higher bandwidth than rectangular guides used at the same operating frequency. Guided mode parameters vary slowly with frequency and there are no low frequency cutoffs for the dominant quasi-TEM mode.

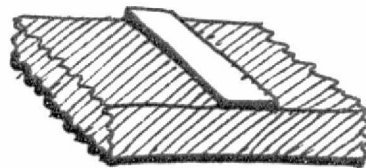
These guides and their properties and uses in the millimeter frequency range are discussed in the following sections.

### 3.2 Materials

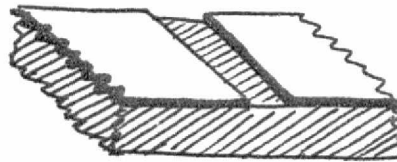
The necessity of support structures for the conductors in planar guides increases the number of criteria to be considered



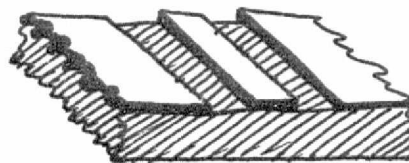
(a) Stripline



(b) Microstrip



(c) Slotline



(d) Coplanar Guide

Figure 3.1 Cross Sections of Various Planar Waveguides



in designing optimum configurations. The usual method of supporting the metal strips and planes is by bonding to dielectric sheets. The choice of dielectric depends on many factors summarized in Table 3.1 [16]. The conductors are usually bonded to one or both sides, as required, of a dielectric layer of the desired thickness through heat and pressure or through a thin layer of a compatible adhesive. Alternatively, the conductors can be formed from an applied layer of conducting paint, with no additional loss.

Circuits are formed from the conductor-clad dielectric by selective removal of the metal depending on the guiding structure being used. Since the tolerances are much more critical at millimeter wavelengths conventional etching techniques can be used only up to around 100 GHz where line widths of a fraction of a millimeter are required. At higher frequencies line widths become so small that ion-beam lithography must be used.

For some guides it is necessary that the dielectric layer be very thin to reduce the conductor spacing so as to obtain the required impedance at millimeter wavelengths. In this case, the physical strength of the dielectric is important for the construction of durable components. If the thermal expansion of the dielectric is much different than that of the conducting layer, the thin components will warp under the heat generated by high power use and performance will deteriorate due to the change in geometry.

TABLE 3.1STRIPLINE CIRCUIT DESIGN

1. Dielectric constant and its variation with frequency and temperature.
2. Dissipation factor and its variation with frequency and temperature.
3. Homogeneity, uniformity and isotropy.
4. Useful temperature range.
5. Dimensional stability with:
  - (a) temperature
  - (b) processing
  - (c) aging
  - (d) humidity
  - (4) cold flow
6. Resistance to chemicals and water.
7. Physical factors including:
  - (a) tensile strength
  - (b) structural strength
  - (c) impact resistance
  - (d) flexibility
  - (e) machinability
  - (f) thermal conductivity
8. Characteristics when clad including:
  - (a) metals available for cladding
  - (b) bond strength
  - (c) need for adhesives for bonding
  - (d) thermal expansion relations
  - (e) blister resistance
  - (f) sizes available (both thickness and area)
9. Processing restraints
  - (a) chemical
  - (b) mechanical.

At millimeter wave wavelengths the inhomogeneities and the anisotropies of some dielectrics due to their weave, particulate nature, or grain, may occur over a significant portion of a wavelength thus making them unsuitable for this use. Fiberglass reinforced compounds are just such dielectrics and are not suitable.

Most glasses, waxes and ceramics are homogeneous and isotropic and can be used for millimeter wave frequency construction.

For many applications at millimeter wave frequencies a high dielectric constant material is required, since a higher field concentration occurs within the dielectric. Field strength decays rapidly into the external material, usually air, and thus avoids spurious coupling to other circuit elements or to a protective or electrically shielding housing. Materials of this type are available such as alumina, some glasses, and semiconductor materials, which have other special applications.

Dielectrics should have stable machining properties. A smooth dielectric surface is a prerequisite to a smooth conducting surface when a metal is bonded to the dielectric. A smooth conducting surface avoids the losses associated with an increase in the surface resistance due to increased conduction current path length. Plastic cold flow during pressing or planing can occur locally in some materials leading to inhomogeneity.

The physical properties of materials suitable for use at millimeter frequencies are given in Table 3.2 [17]. The choice

MATERIAL	DIELECTRIC CONSTANT	LOSS FACTOR	USEFUL TEMPERATURE RANGE (°C)	FLEXIBILITY	COEFF. OF THERMAL EXPANSION ( $\times 10^{-6}$ /°C)	SURFACE FINISH (MICRONS)
Woven TFG	2.55	.0015	- 60° +200°	Good	18.5	N/A
Microfiber TFG (Duroid 5870)	2.33	.0005	- 60° +200°	Good	5	N/A
Microfiber TFG (Duroid 5880)	2.2	.0006	- 60° +200°	Good	32	N/A
Polystyrene	2.53	.0003	- 60° +100°	Very Poor	7	N/A
Reinforced Polystyrene	2.62	.002	- 60° +100°	Poor	5.7	N/A
Polyphenylene Oxide (PPO)	2.55	.0016	- 60° +200°	Fair	29	N/A
Polyolefin	2.32	.00015	- 60° +100°	Excel- lent	4.4	N/A
Quartz Teflon	2.47	.0006	- 60° +200°	Good	18.5	N/A
Polymide (Micaply 5032)	4.8	.01	- 60° +250°	Good	9	N/A
Epsilam-10	10.0	.002	- 60° +150°	Good	11 to 23	N/A
99.5% Alumina	9.9	.00008	up to 500°C	Very Poor	7.5	< 3
Quartz (Fused Silica)	3.78	.0001	up to 500°C	Very Poor	0.55	< 1
Sapphire	9.4 11.6	.00008	up to 500°C	Very Poor	7.7 8.3	< 1
99.5% BeO	6.6	.0004	up to 500°C	Very Poor	7.8	3 to 7
Boron Nitride	4.4	.0003	up to 500°C	Very Poor	1 2	N/A

TABLE 3.2 Parameters for Some Common Substrate Materials

(Reproduced from Howe, Harlan, Jr. "Dielectric Material Development." Microwave Journal, Vol. 21, No. 11, (Nov. 1978), pp. 39-40).

of material is dependent on circuit application and semiconductor substrates are of interest in the construction of millimeter wave integrated circuits. Ceramic materials, such as alumina, have high dielectric constants and have an advantage in that extremely pure homogeneous substrates can be machined to the exact dimensional tolerances required at millimeter wave frequencies.

### 3.3 Stripline and Microstrip

Stripline guides are an extension, into planar form, of coaxial transmission line as shown in Fig. 3.2. Microstrip is a variation of stripline, where one ground plane is removed.

The design of these guides is based primarily on strip width  $w$  and ground plate spacing  $d$ . Plate spacing is chosen to be smaller than one-half wavelength so that the basic TEM mode cannot couple with propagating parallel plate modes which could carry signal power away from the guide and into free space or to other circuits lying in the guide plane. When the plate spacing is less than half of a wavelength, the fields extending outside the strip region are attenuated with an attenuation constant which increases with further reduction of  $d$ . Therefore to achieve the best confinement of guided mode fields, the plate spacing is kept at the smallest possible practical value of around  $1/10$  of a wavelength.

The width of the strip is also kept smaller than one half wavelength in this case to eliminate higher order modes that have

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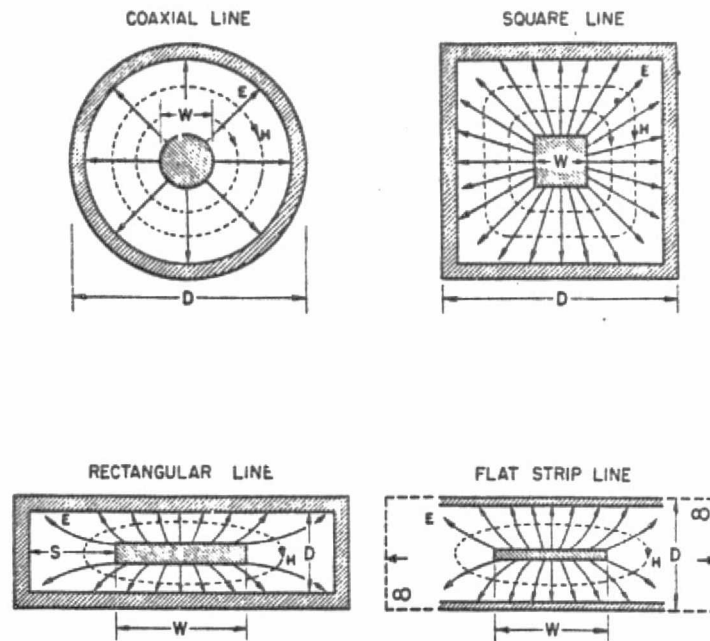


Figure 3.2 Coaxial Line to Stripline Evolution

circumferential field variations. Since these modes have a phase velocity different from that of the dominant TEM mode, excitation of these modes leads to signal distortion.

The impedance of striplines is a complicated expression which depends on the ratios of strip width to ground plate spacing ( $w/d$ ) and strip thickness to ground plate spacing ( $t/d$ ). An approximation for this important guide parameter can be easily found by assuming that there are no field variations in the direction of the guide's width. This approximation is valid for guides which have a large  $w/d$  ratio and small  $t/d$  ratio and this is the usual case for the guides commonly used. In this case

$$Z_c = \frac{1}{4} \sqrt{\frac{\mu}{\epsilon}} \frac{d}{w} \quad (3.1)$$

so that for air filled lines with  $50\Omega$  impedance  $w/d \sim 2$ . The introduction of a dielectric between the strip and ground plates decreases this ratio.

The impedance for microstrip is twice that expressed in 3.1 for strip-ground plate spacing equal to  $d/2$ .

Attenuation can be calculated assuming a uniform TEM-mode in microstrip and stripline. For stripline this attenuation is given by

$$\alpha = \frac{1}{Z_c W \sigma \delta_s} + \frac{2\pi f \epsilon'' W Z_c}{d} \frac{\text{nepers}}{\text{m}} \quad (3.2)$$

where  $Z_c$  is the characteristic impedance,  $\delta_s$  is the skin depth of the metal,  $\sigma$  is the conductivity of the metal and  $\epsilon''$  is the imaginary part of the dielectric constant of the substrate. For microstrip, Eq. (3.2) gives the correct attenuation if the strip-ground plate spacing is  $d/2$  and the impedance,  $Z_c$ , for microstrip is used. Actual loss is higher because there is high current and field concentration near the edge of the strip.

The attenuation and dispersion expressed in terms of actual guide wavelength is given in Table 3.3 [18] for 50 $\Omega$  microstrip line at several millimeter wave frequencies. Microstrip with this characteristic impedance is the most commonly used.

Microstrip, unlike stripline, suffers from radiation into free space. As plate spacing is decreased, the radiation losses are reduced [19] and this loss is kept small at the expense of thin substrate layers; and so, difficulty in manufacturing and reduced power handling capability.

As shown in the table, 50 $\Omega$  microstrip could be expected to have losses of about 28 dB/m at 90 GHz compared to about 3 dB/m for closed rectangular guide at the same frequency.

With the above considerations in mind, most guides have been designed with a strip width on the order of  $1/10$  of a wavelength.

These values are not critical at microwave frequencies and due to the TEM mode of propagation, they are applicable for a



TABLE 3.3

## Parameters of Microstrip Guide

Description	Frequency (GHz)	$\lambda_g$ (cm)	W (cm)	d (cm)	$\alpha$ (dB/cm)	$\alpha$ (dB/ $\lambda_g$ )	$Q_u$
Microstrip-Gold	15	1.210	0.108	0.054	0.0200	0.0242	1124
on Fused Quartz	30	0.605	0.054	0.027	0.0562	0.0340	800
(50 ohm)	60	0.302	0.027	0.014	0.1542	0.0466	573
	90	0.201	0.018	0.009	0.2802	0.563	483

(Reproduced from Knox, Robert M. "Dielectric Waveguide Microwave Integrated Circuits - An Overview." IEEE Trans. MTT, Vol. MTT-24, No. 11 (Nov.1976), pp. 806-14).

wide band of operational frequency use.

At millimeter wave frequencies, construction of guides meeting the above criterion is difficult and costly. The narrow conductor spacing limits the power handling capability due to the generally high field strengths leading to arcing and the high field concentrations near the conductor edges leading to thermal loss.

The small strip widths at millimeter frequencies makes the integration of solid state devices in circuit applications more difficult since such devices would represent a large discontinuity in the guide path and elaborate matching would be required for in-line mounting. An advantage of microstrip and stripline in this application is the potential arrangement for DC biasing of active devices between the strip and ground plane.

Stripline and microstrip circuits are usually designed by approximating lumped element circuits with the desired characteristics by short sections of guide which present the appropriate impedances to the circuit signal. Design parameters for common filters and matching elements are well documented [16]. Most circuits are formed in the strip plane and the construction is simple. Circuits can be mass-produced by photographic techniques. The extension into millimeter wave frequencies of these techniques such that comparable performance results depends on precision processing in this stage of circuit construction.

Microstrip can be constructed with larger dimensions,

compared to a wavelength, despite the considerations mentioned above. This has been called oversized microstrip and has been used in hybrid structures with rectangular waveguide in the millimeter wave frequency range [20].

In such a structure the strip-ground plate spacing is deliberately made to be equal to one quarter of the guided wavelength so that the strip acts as a radiator and couples energy from the  $TE_{10}$  rectangular waveguide mode to a matched load such as a solid state device as shown in Fig. 3.3. Overmoding of the stripline is avoided since these modes would be cut off in the standard rectangular waveguide.

### 3.4 Coplanar Waveguides

Coplanar waveguide utilizes the materials and construction techniques of microstrip and retains the advantages of lightweight, small scale, planar circuit structures.

Additional advantages of using coplanar guide make it an attractive alternative to microstrip. Since coplanar guide has all conducting elements in one plane, mounting of shunt elements between the strip and ground is much easier than with microstrip. Coplanar guide substrates can be much thicker than microstrip substrates and this reduces their dimensional limitations at millimeter wave frequencies.

There is also interest in coplanar waveguide for use in monolithic microwave integrated circuits built on semiconducting

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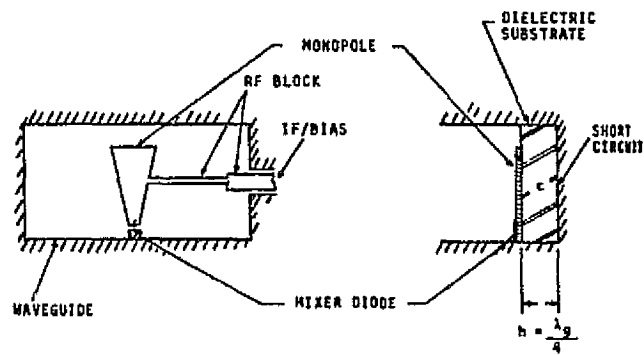


Figure 3.3 Circuit Using Oversized Microstrip

substrates or ferromagnetic semiconductors [21]. If the dielectric constant of the substrate is high enough, a circularly polarized magnetic field occurs with the plane of polarization perpendicular to the surface of the substrate. This makes coplanar guide useful in the construction of non-reciprocal gyro-magnetic devices directly on the substrate surface.

Coplanar guides require a ground plate spacing small compared to one half of a wavelength to avoid radiation losses and so line widths become very small at millimeter wave frequencies as with microstrip. The absence of a ground plane on the lower surface of the substrate limits the heat dissipation for high power applications, since this plane is often used for a heat sink. Coplanar guides which have this metallization layer have a restricted range of impedance values, usually from around 25-80 $\Omega$  [21].

Coplanar guides have a higher attenuation than microstrip or slotline due to the high field and conduction current concentration at the coupled edges of the conductors [21].

### 3.5 Slotline

Slotline is another alternative to microstrip. As mentioned earlier, slotline does not support a TEM or quasi-TEM mode of propagation. The field configuration of this guide consists of electric field lines bridging the slot and magnetic

field lines that form a closed loop perpendicular to the guide plane along every quarter of a guide wavelength. This elliptically polarized magnetic field makes slotline applicable to the construction of non-reciprocal gyromagnetic devices.

The characteristic impedance for slotline can be much higher than that of microstrip, as high as  $200\Omega$  for wide dielectric spacings, and a  $50\Omega$  line is wider than comparable microstrip line thus reducing the adverse effects of miniaturization.

Ohmic losses in slotline are similar to that of microstrip.

## CHAPTER 4

### DIELECTRIC WAVEGUIDES

#### 4.1 Introduction

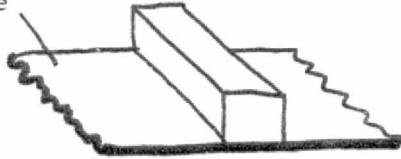
To avoid ohmic losses incurred by confining a transmitted signal inside of a metal guide the signal can be confined by total internal reflection within a system of dielectric interfaces. This is the basis for a broad group of guides known as dielectric waveguides.

To take advantage of planar construction techniques and to provide the ohmic contact required by solid state devices used at millimeter wavelengths these guides have the form of the guides illustrated in Fig. 4.1.

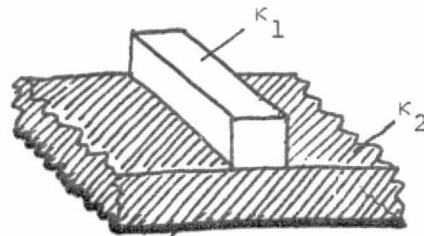
The guides of this type have evolved in the following way. Image line suffers from high ohmic losses due to a high field concentration near the metal for certain modes, thus defeating to an extent the advantages of dielectric guides. Also propagation effects are closely tied to exact dielectric dimensions so that discontinuities in the dielectric can result in radiation loss.

To overcome these disadvantages, a substrate layer is introduced between the guiding strip and the grounding plate. The substrate has a dielectric constant of  $\kappa_2$  and the rib has a dielectric constant of  $\kappa_1$ .

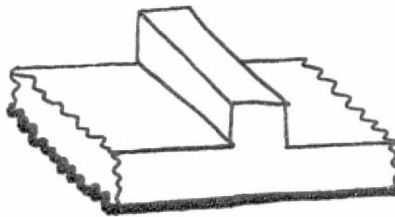
Metal  
Ground  
Plane



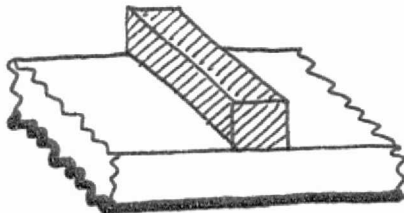
(a) Image Guide



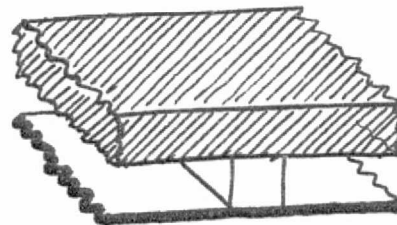
(b) Strip Guide



(c) Ridge Waveguide



(d) Insulated Image Guide



(e) Inverted Strip Waveguide

Figure 4.1 Cross Sections of Various Dielectric Waveguides



When  $\kappa_1 > \kappa_2$  the guide is called the insulated image guide.

Most of the guided energy is in the ridge of the guide and the result is a reduced ohmic loss, but the propagation is still affected strongly by the exact dimensions of the rib. A strip guide, for which  $\kappa_1 < \kappa_2$ , the effects are the opposite, that is, increased stability with respect to guide geometry at the expense of increased ohmic loss. The ridge guide is a special case of these guides, where  $\kappa_1 = \kappa_2$ .

A guide that accomplishes both of the positive aspects of these structures is the inverted strip guide. Energy is carried in a high dielectric constant layer held above the ground plate by a low dielectric constant strip which confines the propagating wave laterally.

Some of the guides that utilize a dielectric plane suffer from leakage of energy since these planes can support a surface mode propagating in any direction. This effect increases losses and can degrade signals travelling in coplanar lines due to coupling of the leaky modes.

In addition to reduced ohmic losses dielectric guides have larger cross-sectional dimensions compared to microstrip. For instance, at 90 GHz a typical 50 $\Omega$  microstrip width would be 0.18mm. An insulated image guide would be about 1 mm wide.

This increase in cross-sectional dimensions is actually an advantage when the miniaturization required at high frequencies become costly to achieve.

Construction of dielectric guides is relatively easy since dielectric materials are often available in tape form and can easily be fastened to the supporting substrate in any planar configuration. For thicker dielectric lines, circuits can be cast and drawn through metal dies to create a uniform cross section.

Inverted strip guide has proven to be very tolerant of "loose" construction techniques and experiments in which passive components formed from strips of Teflon mechanically clamped between the ground plate and the high-dielectric constant quartz upper plate have performed close to theoretical predictions [22].

A major disadvantage to overcome is the integrating of solid state devices to dielectric guides. This is difficult since there is only one conductor available to make ohmic contact.

Radiation from discontinuities is another disadvantage and this can be reduced by careful selection of guide dimensions and metal shielding.

Dielectric guides will become more useful in millimeter-wave circuits as more devices utilizing their unique properties become

available. Recently published papers [23,24], which present detailed theoretical methods for determining the characteristics of these guides will help this process as dielectric guides become as well-known and well-documented as closed rectangular guides and microstrip guides.

The theoretical methods for analyzing these guides is therefore presented in the next two sections. The mode-matching technique shown follows the analysis in [23] although the form of the equations is in terms of the electromagnetic fields rather than the equivalent voltages and currents so that the final determinant equation can form a clearer picture of the propagation mechanism in terms of the field configuration.

#### 4.2 The Mode Matching Technique Used in the Analysis of Dielectric Guides

Dielectric guides are usually analyzed by dividing the transverse geometry into the largest sections which are homogeneous in the  $x$  direction (Fig. 4.2). Fields in each region are described in terms of the modes that can exist on the surface formed by extending the geometry of that region infinitely in the  $\pm x$  direction.

This set of modes consist of a finite number of surface modes and a continuous spectrum of non-surface modes. Analysis can be simplified by introducing a perfectly conducting plate above the

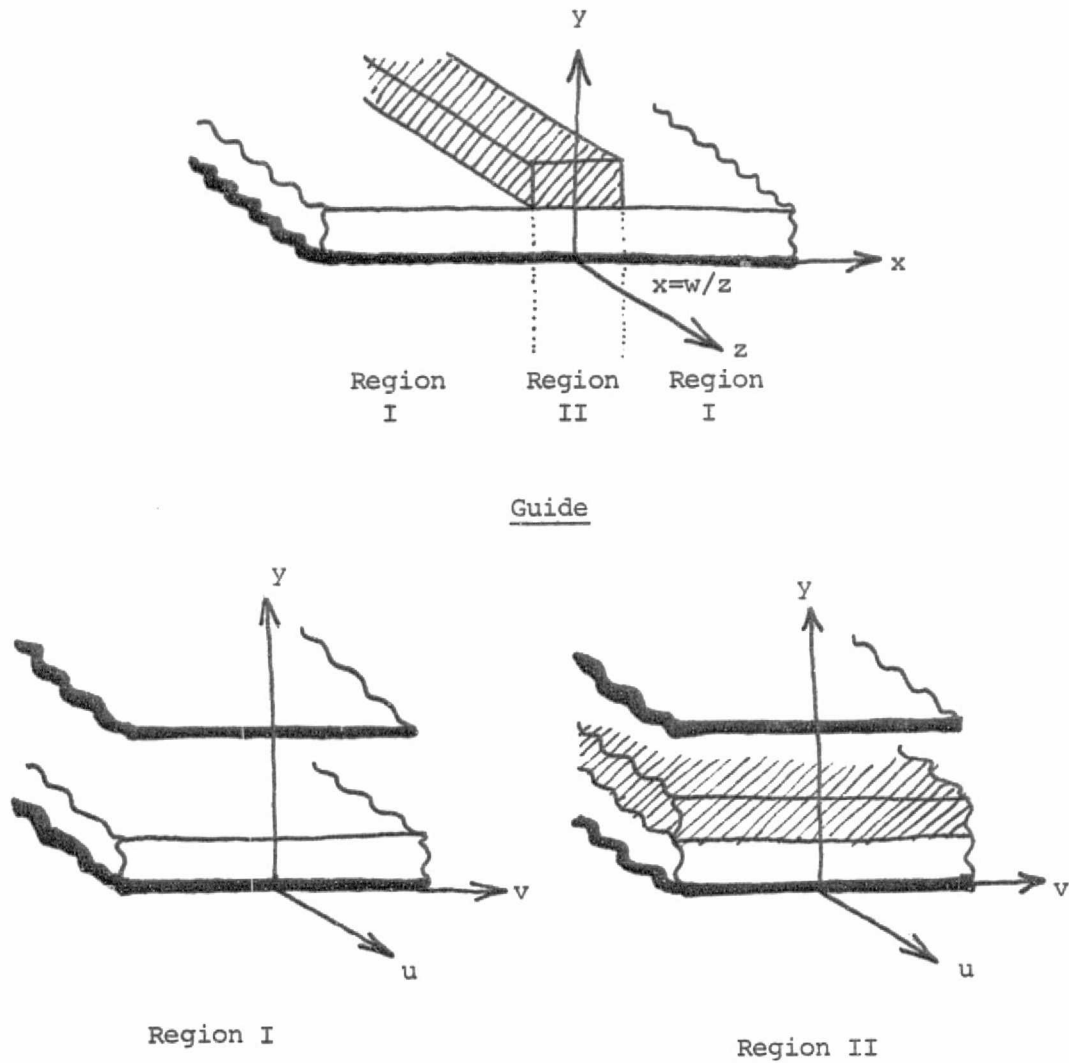


Figure 4.2 The Analysis of a Dielectric Guide in Terms of Uniform Planar Regions.

planar dielectric layer at a distance large enough so that it does not significantly effect the surface mode characteristics. This converts the continuous spectrum of non-surface modes into a discrete spectrum of an infinite number of higher order modes. Therefore, this analysis will describe the fields in each region as an infinite sum of the modes that exist in a dielectric-loaded parallel-plate waveguide.

The direction of propagation of these modes in the infinite parallel-plate guide is an arbitrary direction parallel to the plates and is chosen as the u-axis in a coordinate system, v-y-u, which is rotated about y with respect to the coordinate system of the dielectric guide by an angle  $\theta$  given by

$$\begin{aligned} u &= x \cos \theta + z \sin \theta \\ v &= -x \sin \theta + z \cos \theta \end{aligned} \quad (4.1)$$

Both TE and TM modes can exist between the plates and are required to completely describe propagation in the dielectric guide. The fields of these modes are completely described by the scalar functions  $\phi'$  and  $\phi''$ , and by the propagation wavenumbers  $k_u'$  and  $k_u''$ , where the prime denotes TE modes and the double prime denotes TM modes. The  $\phi$  functions are solutions to the one-dimensional Helmholtz equation.

$$\frac{d^2 \phi}{dy^2} + k_c^2 \phi = 0 \quad (4.2)$$

The wavenumber  $k_c$  is different in each dielectric layer between the plates and is given by

$$k_c = (k_o^2 \kappa_m - k_u^2)^{\frac{1}{2}} \quad (4.3)$$

where  $\kappa_m$  is the dielectric constant of the layer. The quantities  $\phi$  and  $k_u$  can be found by matching the tangential fields at each dielectric or conductor interface. The complete fields in terms of  $\phi$  and  $k_u$  are

#### TE modes

$$E_v' = \phi'(y) \exp(-jk_u' u) \quad (4.4a)$$

$$E_y' = \frac{k_u'}{\omega \mu_o} \phi'(y) \exp(-jk_u' u) \quad (4.4b)$$

$$E_u' = \frac{1}{j\omega \mu_o} \frac{d}{dy} \phi'(y) \exp(-jk_u' u) \quad (4.4c)$$

#### TM modes

$$H_v'' = \phi''(y) \exp(-jk_u'' u) \quad (4.5a)$$

$$E_y'' = \frac{k_u''}{\omega \epsilon_o \kappa(y)} \phi''(y) \exp(-jk_u'' u) \quad (4.5b)$$

$$E_u'' = \frac{1}{j\omega \epsilon_o \kappa(y)} \frac{d}{dy} \phi''(y) \exp(-jk_u'' u) \quad (4.5c)$$

Solution for the above conditions yield an infinite number

of  $\phi$  and  $k_u$  which can be denoted as  $\phi_n$  and  $k_{un}$ .

An alternate formulation of the same problem is the transverse resonant technique [23] which yields the same wavenumbers and mode functions, but expresses the field components in terms of equivalent transmission line voltages and currents along a network that depends on the geometry of the guide in the  $y$  direction. This method allows a complex electromagnetic field problem to be viewed in terms of the more familiar transmission line problem.

Returning to the dielectric guiding structure, it can be seen that the total fields in the guide must be a sum of the fields that can exist in the infinite planar regions. Each parallel plate mode is travelling at an angle,  $\theta_n$  for the  $n$ th mode, which is chosen so that the tangent fields across the boundary,  $x = \pm w/2$ , are continuous (Fig. 4.3). The fields in the parallel plate can be written in terms of the  $x$ - $y$ - $z$  coordinate system of the guide by rotation through  $\theta$ .

The fields will have five non-zero components in general and TE modes become LSE modes and TM modes become LSM modes. These are the following:

LSE modes

$$E_x' = \sin\theta' \exp(-jk_x' x) \phi'(y) \quad (4.6a)$$

$$E_y' = 0 \quad (4.6b)$$

$$E_z' = -\cos\theta' \exp(-jk_x' x) \phi'(y) \quad (4.6c)$$

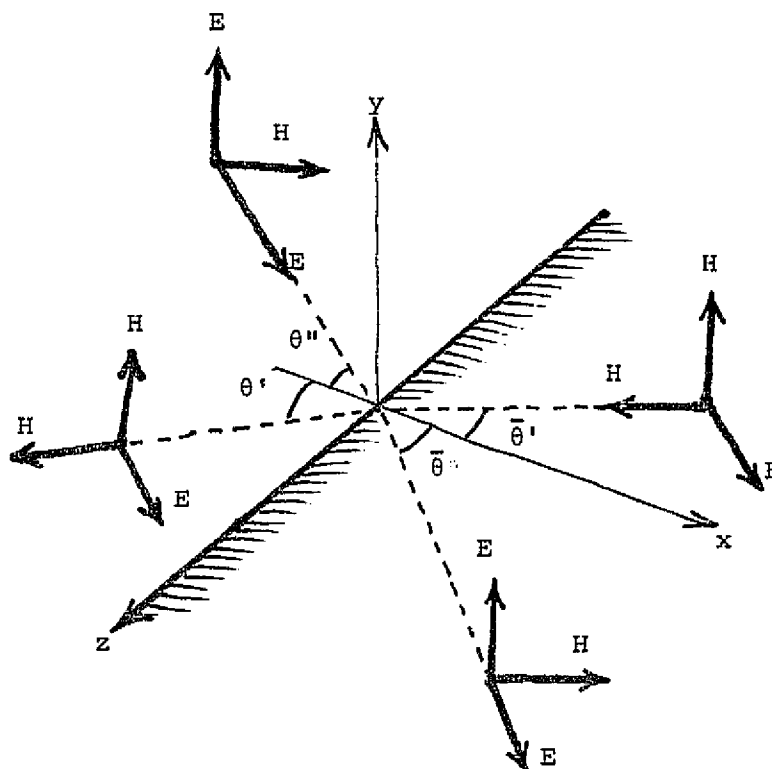


Figure 4.3 Field Components Due to Scattering at  
the Boundary Between Regions.



$$H_x' = \frac{\cos\theta'}{j\omega\mu_0} \exp(-jk_x' x) \frac{d\phi'(y)}{dy} \quad (4.6d)$$

$$H_y' = \frac{k_u'}{\omega\mu_0} \exp(-jk_x' x) \phi'(y) \quad (4.6e)$$

$$H_z' = \frac{\sin\theta'}{j\omega\mu_0} \exp(-jk_x' x) \frac{d\phi'(y)}{dy} \quad (4.6f)$$

#### LSM modes

$$E_x'' = \frac{\cos\theta''}{j\omega\epsilon_0\kappa(y)} \exp(-jk_x'' x) \frac{d}{dy} \phi''(y) \quad (4.7a)$$

$$E_y'' = \frac{k_u''}{\omega\epsilon_0\kappa(y)} \exp(-jk_x'' x) \phi''(y) \quad (4.7b)$$

$$E_z'' = \frac{\sin\theta''}{j\omega\epsilon_0\kappa(y)} \exp(-jk_x'' x) \phi''(y) \quad (4.7c)$$

$$H_x'' = -\sin\theta'' \exp(-jk_x'' x) \phi''(y) \quad (4.7d)$$

$$H_y'' = 0 \quad (4.7e)$$

$$H_z'' = \cos\theta'' \exp(-jk_x'' x) \phi''(y) \quad (4.7f)$$

The propagation factor  $\exp(-jk_z z)$  is not included in the above expressions since it is known that the fields of all modes of both types in both regions must be in phase along the  $z$  direction. Therefore this factor is common to all field expressions.

It is also noted that

$$\begin{aligned} k_{xn} &= k_{un} \cos \theta_n \\ k_z &= k_{un} \sin \theta_n \end{aligned} \quad (4.8)$$

for the  $n$ th mode of either LSE or LSM types.

The fields can now be matched at  $x = \pm w/2$ . The fields must be summed in the proper proportions so that their tangential components are continuous across the boundary. The following constants of proportion are used:

$A_n$  = amplitude factor of the  $n$ th LSE mode in the inner region at the boundary.

$\bar{A}_n$  = amplitude factor of the  $n$ th LSE mode in the outer region at the boundary.

$B_n$  = amplitude factor of the  $n$ th LSM mode in the inner region at the boundary.

$\bar{B}_n$  = amplitude factor of the  $n$ th LSM mode in the outer region of the boundary.

The amplitude factor of a mode travelling toward the boundary is different from the amplitude factor of the same mode travelling away from the boundary. These two factors will be differentiated by arrows indicating the direction of travel.

The boundary conditions are then given as follows:

$$\sum_n (B_{\rightarrow n} + B_{\leftarrow n}) E''_{yn} = \sum_n (\bar{B}_{\rightarrow n} + \bar{B}_{\leftarrow n}) \bar{E}''_{yn} \quad (4.9a)$$

$$\sum_n (\bar{A}_n + \bar{A}_n) H_{yn}' = \sum_n (\bar{A}_n + \bar{A}_n) \bar{H}_{yn}' \quad (4.9b)$$

$$\begin{aligned} \sum_n [(\bar{B}_n - \bar{B}_n) E_{zn}'' + (\bar{A}_n - \bar{A}_n) E_{zn}'] \\ = \sum_n [(\bar{B}_n - \bar{B}_n) \bar{E}_{zn}'' + (\bar{A}_n - \bar{A}_n) \bar{E}_{zn}'] \end{aligned} \quad (4.9c)$$

$$\begin{aligned} \sum_n [(\bar{B}_n - \bar{B}_n) H_{zn}'' + (\bar{A}_n - \bar{A}_n) H_{zn}'] \\ = \sum_n [(\bar{B}_n - \bar{B}_n) \bar{H}_{zn}'' + (\bar{A}_n - \bar{A}_n) \bar{H}_{zn}'] \end{aligned} \quad (4.9d)$$

The fields  $E_y'', H_y', E_z', E_z'', H_z', H_z''$  are given as before and the overbar indicates these field components of the solution to the outer region.

The sets of scalar functions  $\{\phi_n'(y)\}$  and  $\{\phi_n''(y)\}$  are orthogonal sets and have the following property

$$\int_0^h \phi_n'(y) \phi_m'(y) dy = \delta_{nm} \quad (4.10)$$

$$\int_0^h \phi_n''(y) \phi_m''(y) \frac{1}{\kappa(y)} dy = \delta_{nm}$$

where

$$\delta_{nm} = \begin{cases} 1, & n = m \\ 0, & \text{otherwise} \end{cases}$$

Utilizing this property, the above continuity equations can

be converted into four infinite systems of equation that are linear in terms of the amplitude factors. Each equation can be multiplied by  $\phi_m'$  or  $\phi_m''$  and then integrated over  $y$  from  $y=0$  to  $y=h$  to eliminate the  $y$  dependence in each term and give the resulting linear equations. These equations can be written in matrix form

$$a[\vec{B} + \vec{B}] = \bar{a}[\vec{B} + \vec{B}] \quad (4.11a)$$

$$b[\vec{A} + \vec{A}] = \bar{b}[\vec{A} + \vec{A}] \quad (4.11b)$$

$$c[\vec{B} - \vec{B}] + d[\vec{A} - \vec{A}] = \bar{c}[\vec{B} - \vec{B}] + \bar{d}[\vec{A} - \vec{A}] \quad (4.11c)$$

$$e[\vec{B} - \vec{B}] + f[\vec{A} - \vec{A}] = \bar{e}[\vec{B} - \vec{B}] + \bar{f}[\vec{A} - \vec{A}] \quad (4.11d)$$

where the bracketed quantities are column matrices with the sum or difference of the amplitude factor for the  $n$ th mode in the  $n$ th position. The small letter quantities are matrices characterizing the coupling of modes at the boundary and are given in terms of the field expressions.

Although the calculation is rather long, the amplitude factors of the modes in the outer region can be eliminated through the manipulation of the above matrix equations. The result can be written

$$[\vec{B}] = S_{11}[\vec{B}] + S_{12}[\vec{A}] \quad (4.12a)$$

$$[\vec{A}] = S_{21}[\vec{B}] + S_{22}[\vec{A}] \quad (4.12b)$$

Here again the bracketed quantities are column vectors of the amplitude factors. The S matrices can be written entirely in terms of the matrices  $a, \bar{a}, b, \bar{b}, c, \bar{c}, d, \bar{d}, e$  and  $\bar{e}$ . The above expressions can conveniently be written in terms of super matrices of the form

$$\begin{bmatrix} \vec{B} \\ \vec{A} \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} \vec{B} \\ \vec{A} \end{bmatrix} \quad (4.13)$$

The matrices  $S_{11}$  and  $S_{22}$  are the reflection coefficients of the  $TM_0$  and  $TE_0$  modes respectively, and the matrices  $S_{12}$  and  $S_{21}$  are the mode conversion coefficients for  $TE_0$  to  $TM_0$  to  $TE_0$ , respectively. For a waveguide mode, waves must add in phase after total internal reflection at the boundary at  $x = w/2$  and travelling the width of the center region; that is,

$$\begin{bmatrix} \psi'' & 0 \\ 0 & \psi' \end{bmatrix} \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} B \\ A \end{bmatrix} = \pm \begin{bmatrix} B \\ -A \end{bmatrix} \quad (4.14)$$

where  $\psi'$  and  $\psi''$  are diagonal matrices and

$$\begin{aligned} \psi''_{nn} &= e^{-jk''_{xn} w} \\ \psi'_{nn} &= e^{-jk'_{xn} w} \end{aligned} \quad (4.15)$$

The  $\pm$  sign in the above equation indicate modes of odd and even symmetry. This equation can be rewritten

$$\begin{bmatrix} \psi'' S_{11} \mp I & \psi'' S_{12} \\ \psi' S_{21} & \psi' S_{22} \pm I \end{bmatrix} \begin{bmatrix} B \\ A \end{bmatrix} = 0 \quad (4.16)$$

A nontrivial solution to this equation exists only if the determinantal equation holds:

$$\begin{vmatrix} \psi'' S_{11} \mp I & \psi'' S_{12} \\ \psi' S_{21} & \psi' S_{22} \pm I \end{vmatrix} = 0 \quad (4.17)$$

An examination of the previous expressions leading to the determinantal equation reveals that, given the geometry of the guiding structure, the only undetermined quantity is  $k_z$ , the propagation wavenumber of the guide. Intermediate quantities used can be expressed in terms of  $k_z$ :

$$\sin \theta_n = \frac{k_z}{k_{un}}$$

$$\cos \theta_n = \left[ 1 - \left( \frac{k_z}{k_{un}} \right)^2 \right]^{\frac{1}{2}} \quad (4.18)$$

$$k_{xn} = \left[ k_{un}^2 - k_z^2 \right]^{\frac{1}{2}} .$$

and the  $k_{un}$  have been determined.

In general  $k_z$  will be a complex number  $\beta_z - i\alpha_z$ . Since the materials in the guide have been assumed to be lossless, the imaginary part of the wavenumber represents loss due to propagation

of energy away from the guide. This is a leakage effect and is due to conversion at the boundary of the completely bound mode-type to the oppositely polarized mode-type which is unbounded.

As an alternative to the field analysis of the regions of the guide, the geometry in the  $x$  direction can be represented by a set of transmission lines for each region, one line for each mode that can propagate in that region. The continuity condition on the fields translates into a set of relationships between the mode voltages and currents on the equivalent network. This is the approach used in [23] and the resulting expressions are also put into matrix form, and  $k_z$  is found from a determinantal equation.

Since the matrices in both approaches to analysis are infinite the higher order terms must be discarded for practical numerical solutions.

Calculated values of  $\alpha$  using this method are shown in Fig.

4.4 [23]. Leakage is at a minimum when  $Wk_x$  of the dominant leakage mode is a multiple of  $2\pi$ .

#### 4.3 The Effective Dielectric Constant (EDC) Method Used in the Analysis of Dielectric Guides

As in the mode matching technique a dielectric guide is divided into sections which have the same dielectric geometry along the  $y$  direction. Uniform planar regions with this same geometry support a finite number of surface modes whose properties are

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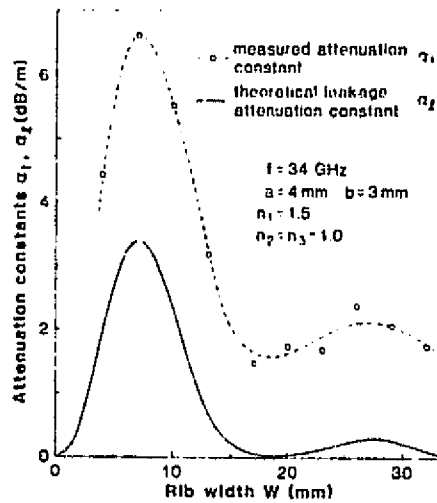


Figure 4.4 Theoretical and Measured Values for  
Attenuation Due to Leakage vs. Strip  
Width.

(Reproduced from Peng, Song-Tsuen, and Oliner, Arthur A.  
 "Guidance and Leakage Properties of a Class of Open  
 Dielectric Waveguides: Part I. Mathematical Formulations."  
IEEE Trans. MTT, Vol. MTT-29, No. 9 (Sept. 1981),  
 pp. 843-55).



then used to determine the characteristics of the entire guided modes of the dielectric guide (Fig. 4.5).

A major assumption used in the EDC technique is that only a single one of these surface modes propagates in each region of the guide. The mode is assumed to be of the same order and type (TE or TM) in each region. These assumptions are reasonable as long as the boundary does not represent an extreme electrical discontinuity since the higher order modes that would be excited there are ignored.

Given the wavenumber of the surface wave propagating in each region which are found by solution to the field problem of the surface guide, an effective dielectric constant,  $\kappa_e$ , is defined as

$$\kappa_e = \left( \frac{k_u}{k_o} \right)^2 \quad (4.19)$$

If each region is completely characterized by the effective dielectric constant of that region, then an effective guiding structure can be formed by representing each region by a dielectric layer having that region's width and effective dielectric constant. The formation of the effective guiding structure is illustrated in Fig. 4.5.

The formation of an effective guide in this manner allows the original waveguide structure to be analyzed as a dielectric slab guide supporting a surface wave mode propagating in the z direction.

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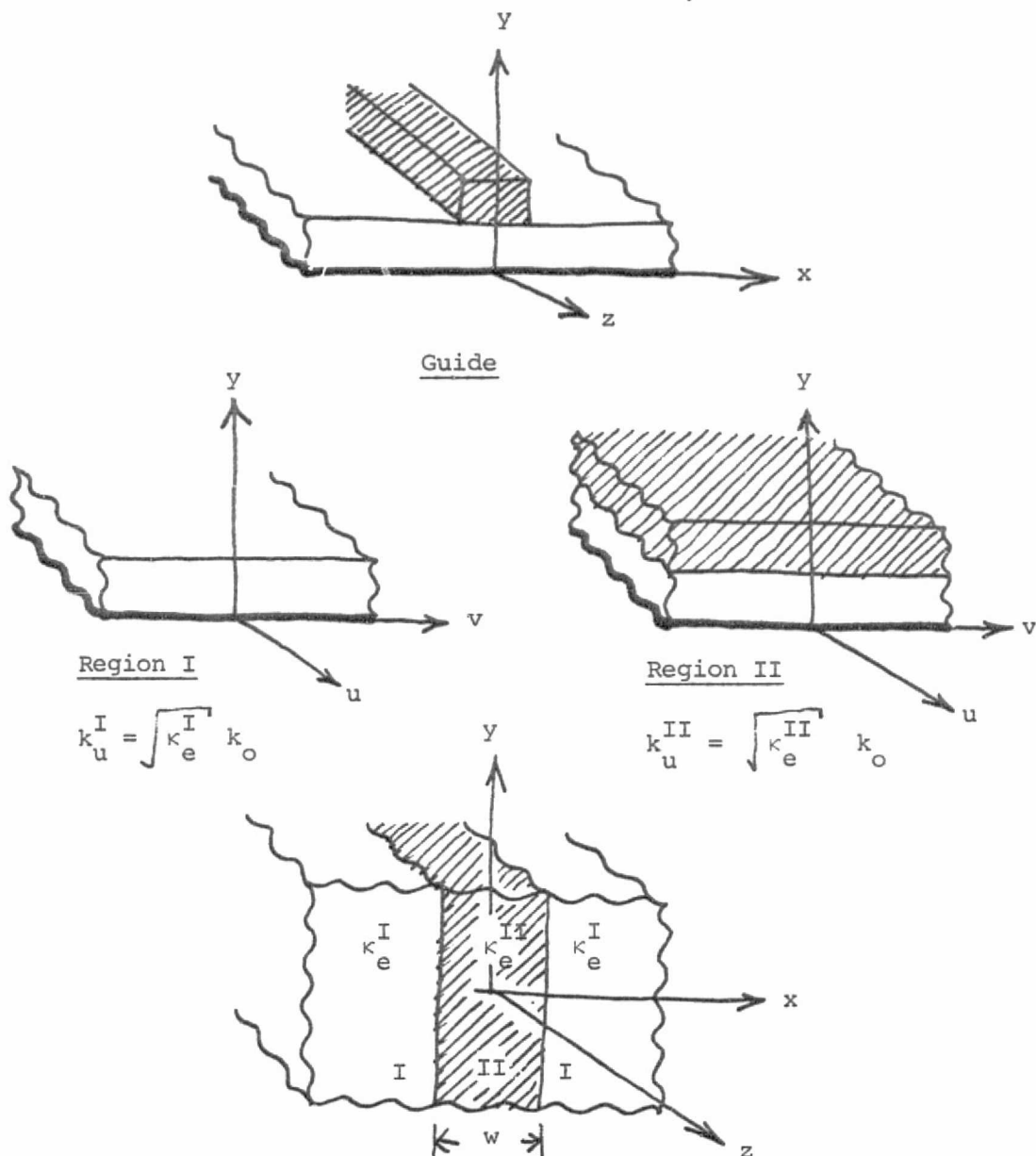


Figure 4.5 The Evolution of an Effective Guiding  
Structure from the Dielectric Guide Regions.

To complete the analysis, the polarization of the fields in the guide must be specified. If the single surface mode in each region is TE then the effective guide structure must support a TM surface wave so that the y component of electric field in each representation is zero, and the field is of the LSE type. Similarly if each region supports a TM surface mode then the effective guide supports a TE surface mode for LSM type guided modes. These relations are illustrated in Figs. 4.6.

In the illustrated examples there are only two regions, I and II. For this case, if the width of region II,  $w$ , is zero, the guide supports a surface wave of the type used to characterize region I and the guide wavenumber is

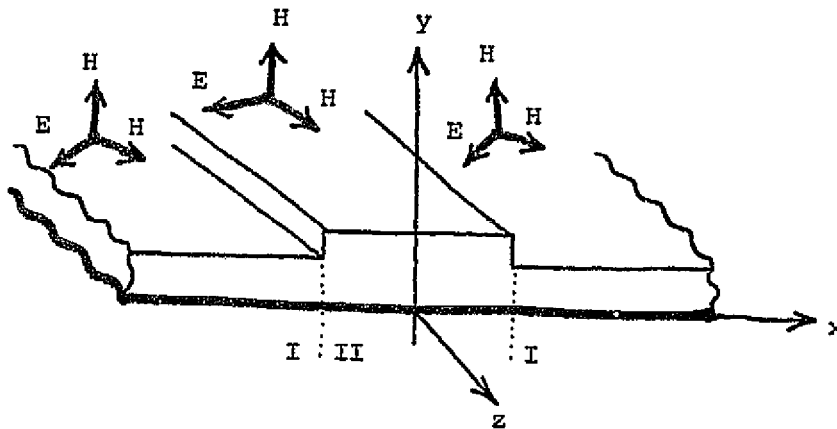
$$k_z = \sqrt{\kappa_e^I} k_0 \quad (4.20)$$

In the other extreme, as  $w$  tends to infinity, the guide supports the surface wave of region II and

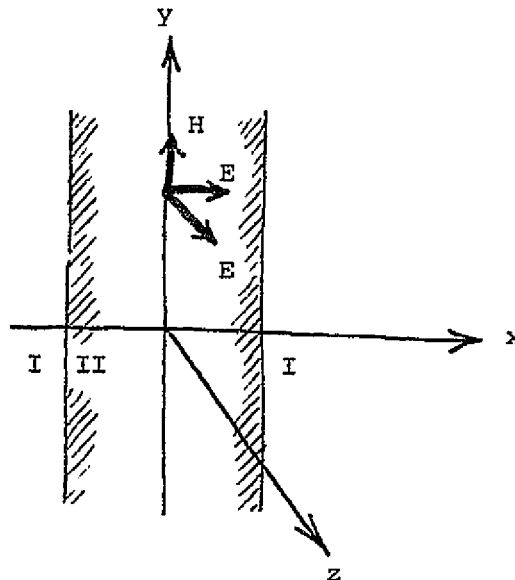
$$k_z = \sqrt{\kappa_e^{II}} k_0 \quad (4.21)$$

For any width  $w$ ,  $k_z$  is bound by

$$\sqrt{\kappa_e^I} k_0 \leq k_z \leq \sqrt{\kappa_e^{II}} k_0 \quad (4.22)$$



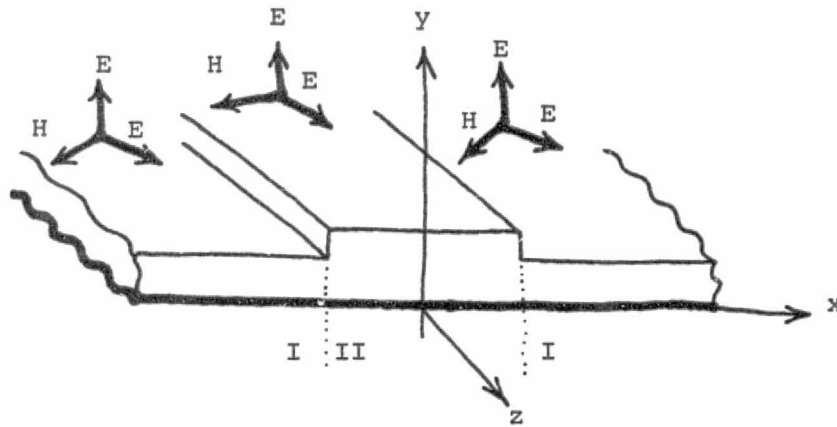
(1) TE surface modes propagate in each region.



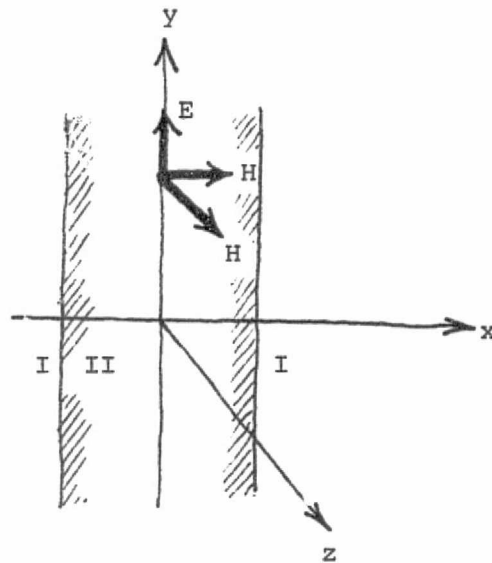
(2) A TM mode propagates in the effective guiding structure.

Figure 4.6 The Two Types of Longitudinal Section Modes.

(a) LSE Mode ( $E_y = 0$ ).



(1) TM surface modes propagate in each region.



(2) A TE mode propagates in the effective guiding structure.

Figure 4.6 The Two Types of Longitudinal Section Modes.

(b) LSM Mode ( $H_y \neq 0$ )

or

$$\kappa_e^I \leq \kappa_g \leq \kappa_e^{II} \quad (4.23)$$

where

$$\kappa_g = \left( \frac{k_z}{k_o} \right)^2 \quad (4.24)$$

It is noted that  $\kappa_e^{II}$  must be greater than  $\kappa_e^I$  for the effective guiding structure to support a surface wave.

As an application of this method of analysis to a practical aspect of guide design, beyond determining the guide wavenumber, it will be shown that  $w$  for a guide can be chosen to avoid the leakage effect shown to exist by the more accurate mode-matching technique of the previous section.

The leakage is due to mode conversion at the boundaries at  $x = \pm w/2$ . If the guide supports a dominant LSE type mode then a dominant TE mode propagates in the inner region II and this mode is cut off in the outer region I. But the inner and outer regions can support TM modes, and if the mode conversion at the boundary excites a TM mode that is above cutoff in the outer region, energy can propagate away from the guide axis in a leakage effect.

To determine whether or not leakage can occur it is assumed that the guide wavelength is unchanged by leakage and mode conversion, and only one surface mode of each type, TE and TM, exists in each region and that these modes are of the same order (such

as  $TE_0, TM_0$ ).

A LSE mode is assumed to propagate in a dielectric ridge guide. A TM type wave is excited in the outer region and this wave travels in a direction such that the projection of its wave vector,  $k_{TM}^{II}$ , along the z axis,  $k_{zTM}^{II}$ , is equal to the wavenumber of the guided LSE mode  $k_g$  so that the fields are in phase along the boundary since

$$(k_{TM}^I)^2 = (k_{xTM}^I)^2 + (k_{zTM}^I)^2 \quad (4.25)$$

it follows that

$$(k_{xTM}^I)^2 = (k_{TM}^I)^2 - (k_g)^2 \quad (4.26)$$

If the wavenumber  $k_{xTM}^{II}$  is real then the TM mode in the outer region has a component of propagation in the direction perpendicular to the guide axis and the guide leaks. That is, leakage occurs for

$$(k_{xTM}^I)^2 > 0 \quad (4.27)$$

or

$$k_{TM}^I > k_g$$

using the equality above. Dividing this inequality by  $k_0$  gives the condition for leakage in terms of effective dielectric constants:

$$\kappa_{TM}^I > \kappa_g \quad (4.28)$$

For a ridge waveguide the only parameter that varies from the inner to outer regions is the dielectric thickness,  $t$ . A dispersion curve can be plotted which shows the effective dielectric constant for several surface modes as a function of normalized dielectric thickness,  $k_0 t$  (Fig. 4.7). By choosing  $k_0 t^I$  and  $k_0 t^{II}$  as the normalized thickness of the outer and inner regions, respectively, the constants  $\kappa_{TE}^I$ ,  $\kappa_{TE}^{II}$ ,  $\kappa_{TM}^I$ ,  $\kappa_{TM}^{II}$  for a particular mode number can be read from the graph. Assuming again that the guided mode is of the LSE type,  $\kappa_g$  can be calculated using  $\kappa_{TE}^I$  and  $\kappa_{TE}^{II}$  and can be shown on the graph by indicating the normalized width of the rib  $wk_0$  on a second ordinate scale. In this way  $wk_0$  and so the rib width  $w$  can be chosen so that leakage does not occur with restrictions as follows. Recall that

$$\kappa_{TE}^I \leq \kappa_g \leq \kappa_{TE}^{II} \quad (4.29)$$

Therefore, if  $\kappa_{TM}^I > \kappa_{TE}^{II}$  then no ridge width is sufficient to eliminate leakage. If  $\kappa_{TM}^I < \kappa_{TE}^I$  then leakage will not occur for any ridge width. For  $\kappa_{TE}^I < \kappa_{TM}^I < \kappa_{TE}^{II}$  leakage will occur for all  $w < w_c$  and will not occur for  $w > w_c$  where  $w_c k_0$  can be determined from the graph (Figs. 4.8).

The same analysis can be applied to LSM type modes in an obvious way.

In a variation to this analysis the dielectric thicknesses  $t_1$  and  $t_2$  can be varied to eliminate leakage.



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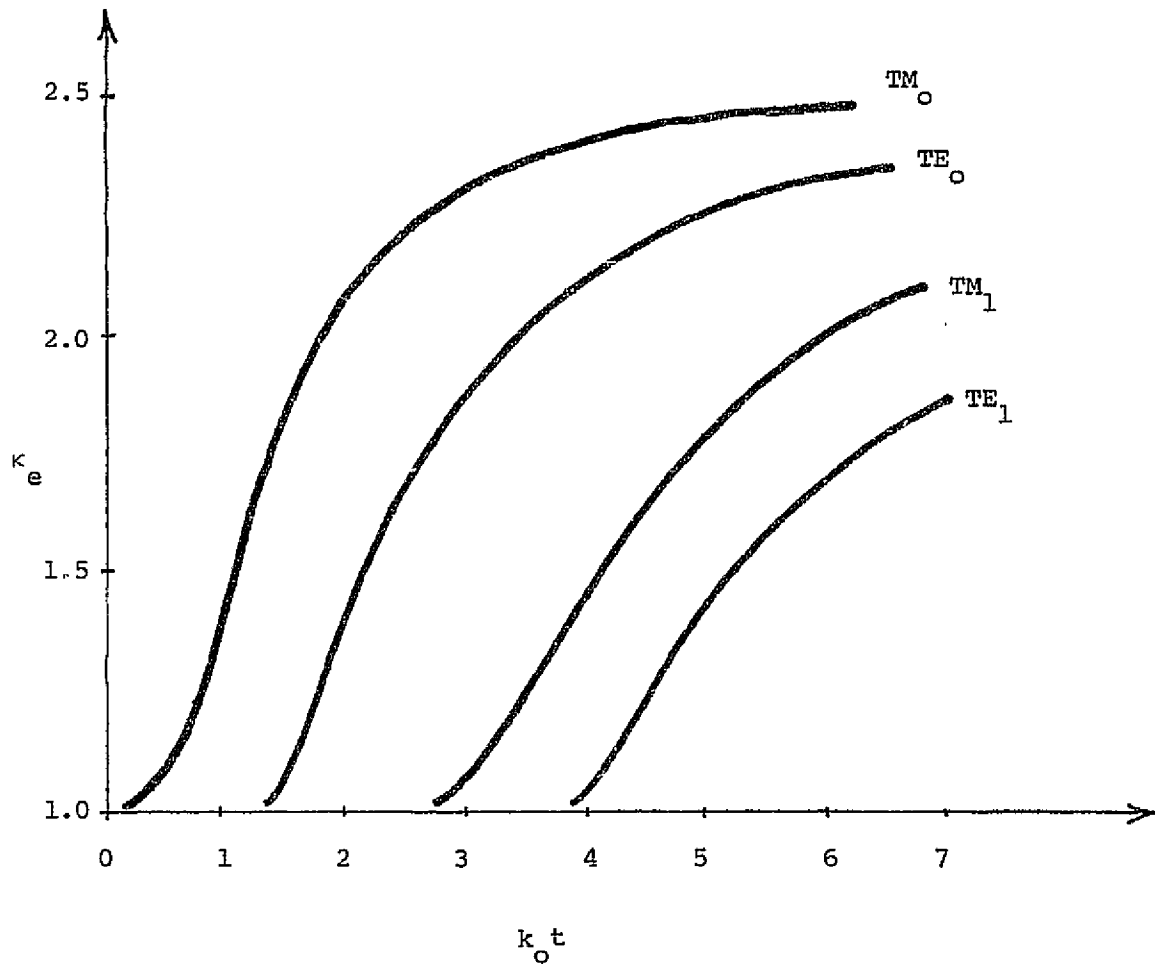


Figure 4.7 Dispersion Curves for Dielectric Slab Guide Modes

( $\kappa_{\text{ridge}} = 2.54$ )

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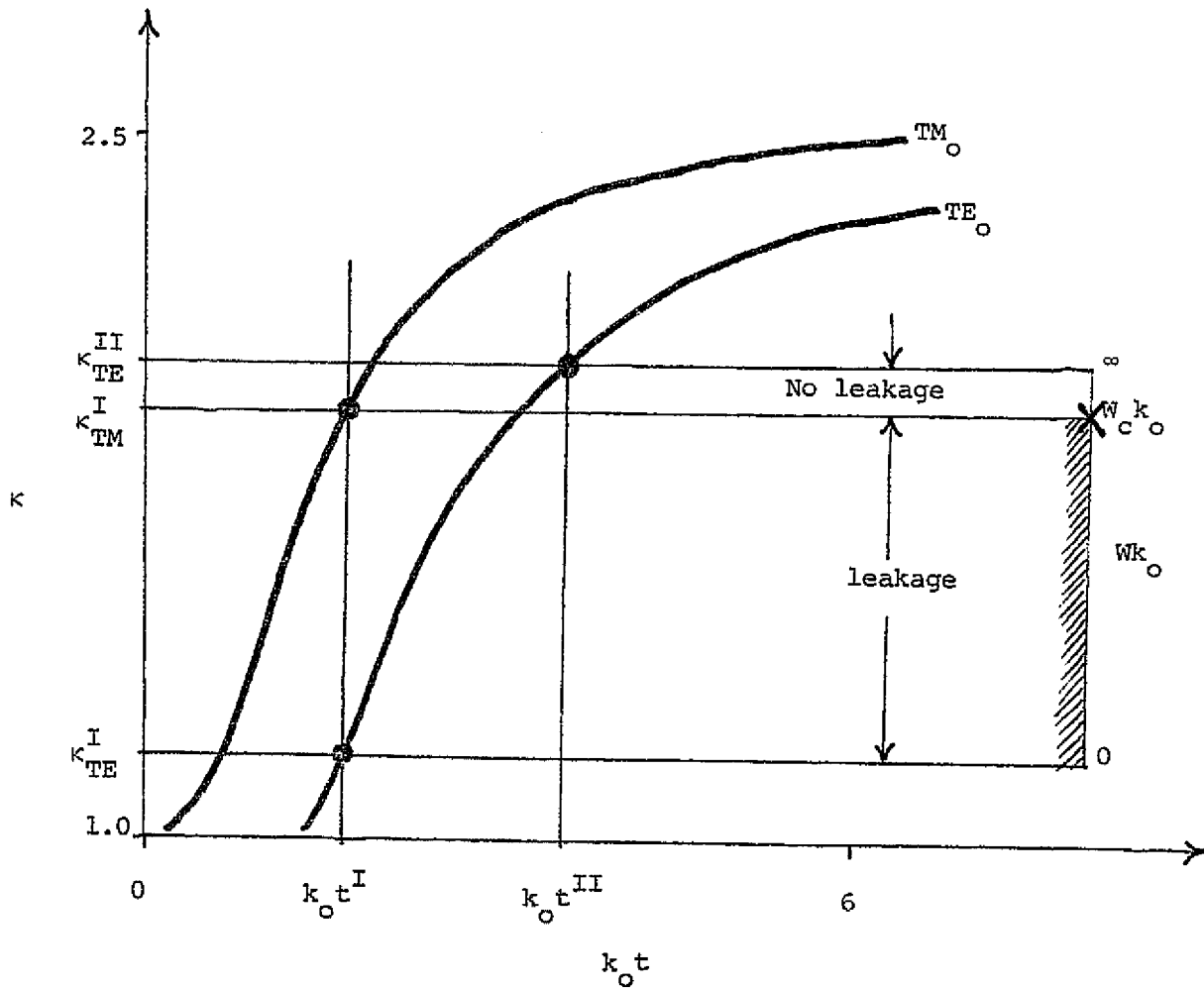


Figure 4.8a Ridge Guide - LSE Mode.

$t^I$  and  $t^{II}$  chosen so that the guide leaks  
for  $W > W_c$ .

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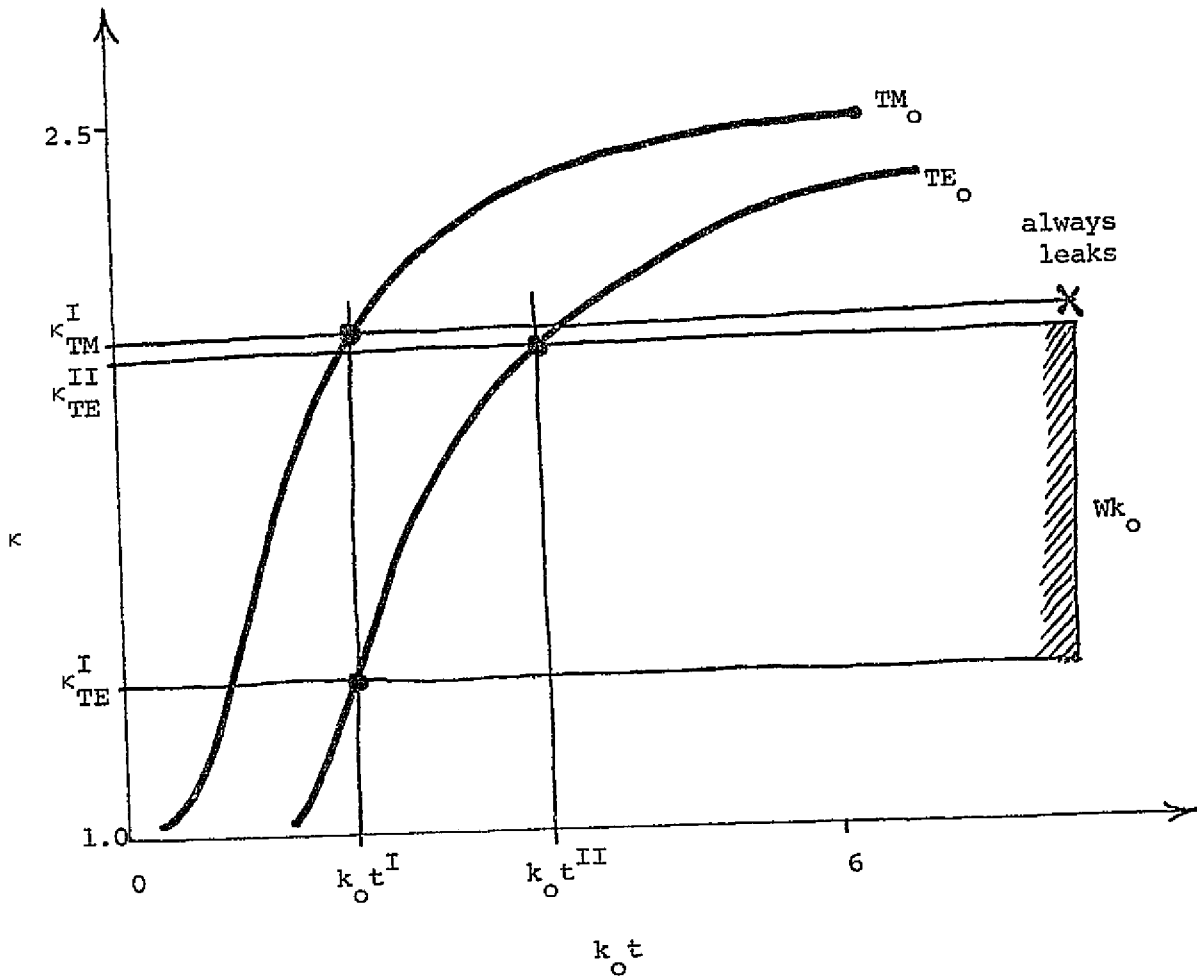


Figure 4.8b Ridge Guide - LSE Mode

$t^I$  and  $t^{II}$  chosen so that the guide always leaks.



## CHAPTER 5

### COMPONENTS AND DEVICES

#### 5.1 Introduction

At frequencies lower than 30 GHz, microwave circuit design has included the use of lumped circuit elements (whose dimensions are less than  $1/10$  of a wavelength) and distributed elements which are formed from transmission line sections. Lumped elements are preferred for broad-band physically small circuit applications. The parasitics which occur in miniature elements lower the  $Q$  factor of such circuits. For narrow-band, high  $Q$  circuits, distributed elements are recommended.

At millimeter wave frequencies, lumped circuit elements are nearly impossible to construct. Even when microfabrication techniques are used; i.e. as to deposit metallic geometries for monolithic microwave integrated circuits, lumped elements are considered to be useful only to about 20 GHz.

Distributed elements are therefore expected to dominate circuit design at millimeter wavelengths. Since most reactive elements are composed of resonant stubs of transmission line sections, bandwidth is limited by the frequencies over which the stubs are resonant. The performance of distributed devices such as couplers depends on guide length and sometimes good performance requires an increase in circuit size, which can be a

disadvantage when small circuit dimensions are required.

All of the waveguide types described in the previous chapters have been used to build passive devices such as couplers, hybrid junctions, resonant filter structures, etc. The quality of these devices are related to the performance of each guide as a transmission line, but additional advantages and disadvantages important to the component design are unique to the guide used. For this reason, many complete circuits and systems are hybrids, to make use of the best guide-type for each component.

In the following sections, devices constructed from the various guide types will be compared to determine the usefulness of these guides in millimeter-wave circuit applications.

## 5.2 Launchers

The ability to excite a propagating mode in a waveguide is of primary importance. The effectiveness of any launching structure used for this purpose is effected by the field distribution in the guide and that of the signal source. Some launching mechanisms present a reactive load to the guide and this must be tuned out with an appropriate matching network.

The dominant  $TE_{10}$  mode in rectangular guides is easily excited by probes fed by coaxial line and by aperture coupling with other rectangular guides. In most cases, though, the millimeter source is fitted with a standard rectangular waveguide fitting so that transitions from this waveguide to others are of more

importance. This is also true since many instruments for millimeter wave measurements are adapted to rectangular guides.

Therefore, transmissions between rectangular guides and other guide types will be examined.

Stripline and microstrip can also be fed from coaxial line, but as described above a transition to rectangular guide is more practical at millimeter wave frequencies. A method for coupling the rectangular guide mode to microstrip line by selecting a substrate thickness equal to one quarter wavelength was described in Section 3.3. This hybrid structure was used to direct signal energy to a solid state device and does not represent the most common method used to couple these two guides. A gradual taper from rectangular guide to double ridged guide allows stripline to be introduced in the gap between the two ridges. This structure is shown in Fig. 5.1 [25]. A similar structure can be used for transitions from rectangular guide to stripline.

Other planar guides such as slotline can be fed from stripline or can be fed from the dielectric and fin elements of finline which are mounted in the E-plane of rectangular guides.

Transition from rectangular guide to finline itself takes the form of tapered dielectric and fin section as shown in Fig. 5.2 [26].

Transitions from rectangular guides to dielectric guides are the most difficult since guided signal energy is not always

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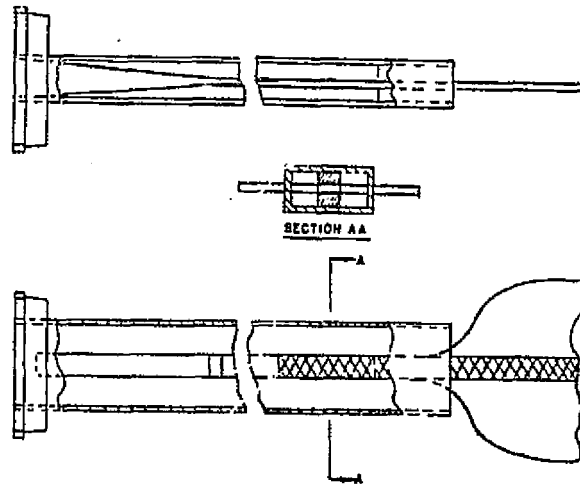


Figure 5.1 A Waveguide to Microstrip Transition.



confined to the immediate vicinity of the guiding strip or layer. In this case energy is coupled to rectangular guide through a horn which extends far enough in the transverse direction to capture most of the signal energy. An example of such a transition used for dielectric image line is shown in Fig. 5.3 [27].

If the impedance of guide sections are known as a function of some easily-varied geometrical parameter, such as line width, tapered transitions of the type required for the above guide transitions can be designed according to design specifications using exponential, Chebyshev or other tapers.

### 5.3 Directional Couplers

Directional couplers represent another way to excite a propagating mode in a waveguide, that is, from a similar guide carrying the desired propagating mode. A directional coupler is a four port device in which power incident on one port is coupled to two other ports but not the fourth. The coupling,  $C$ , of a directional coupler is a measure of the power available at the input as a fraction of the coupled power expressed in decibels. The directivity  $D$ , is a measure of the coupled power as a fraction of the power available at the isolated port.

The guides that are useful at the millimeter wavelength can be used to form distributed, aperture and beam-splitter type couplers as described below.

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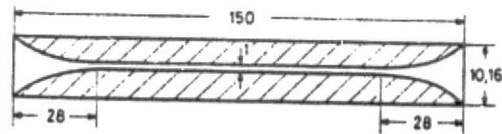


Figure 5.2 A Rectangular Waveguide to Finline Transition

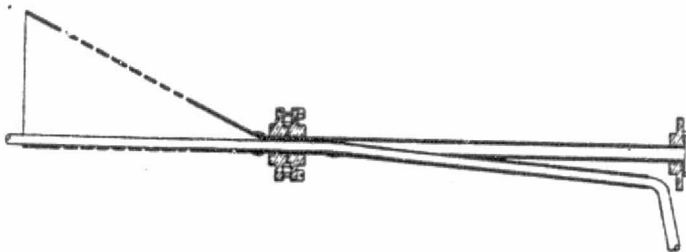


Figure 5.3 A Horn Transition for Dielectric Image  
Guide

When an open guide, usually of the planar type, is placed as a second line in proximity to a primary line directional coupling occurs as a result of the interference of the fringing fields, excited when the primary guide carries signal power. This type of directional coupler is called distributed and the theory of its operation follows.

Figure 5.4 illustrates that, due to the symmetry of the adjacent lines, two types of modes can exist on the coupled lines. These modes are called the even and odd modes, due to the symmetries of their field configurations, and they have propagation constants  $k_{ze}$  and  $k_{zo}$ , respectively.

If signal energy is input to port 1, the field configuration is considered to be a superposition of even and odd modes giving zero power at port 2 of the coupler.

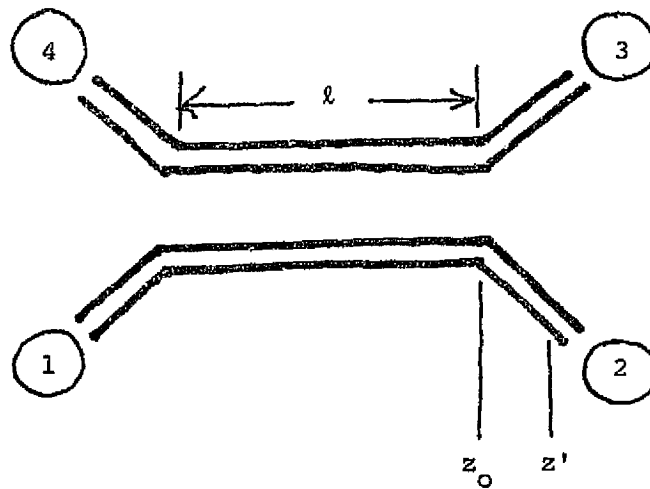
Then, using the propagation constants  $k_{ze}$  and  $k_{zo}$ , the phase of the even and odd modes can be determined at the output of the coupler to determine the output power at each of these two ports.

For a coupler length,  $L$ , given by

$$L = \frac{\pi}{k_{ze} - k_{zo}} \quad (5.1)$$

the two modes will cancel at port 2 and a complete transfer of energy from port 1 to port 3 will occur. For an arbitrary length,  $l$ , the power ratio  $P_3/P_2$  is

C-2



a) Top View Showing Geometry of Distributed  
Directional Coupler



b) Electric Field Lines in a Transverse Section  
of Coupled Lines Showing Symmetry of Modes

Figure 5.4 Distributed Coupler and Coupled  
Line Modes.

$$\frac{P_3}{P_2} = \tan^2 \left( \frac{\pi}{2} \frac{\ell}{L} \right) \quad (5.2)$$

When the coupling between the connecting lines is significant  $k_{ze}$  and  $k_{zo}$  can be considered functions of the transverse geometry, which in turn is a function of  $Z$ . The variable  $\ell$  can then be replaced by  $\ell'$ :

$$\ell' = \ell + \frac{\Delta\phi}{L} L. \quad (5.3)$$

where

$$\Delta\phi = 2 \int_{z_0}^{z'} [k_{ze}(z) - k_{zo}(z)] dz. \quad (5.4)$$

and  $z_0$  is the location of the junction to the connecting lines and  $z'$  is beyond where any significant coupling would still occur. In this case the coupler is considered symmetric about  $z = 0$ .

This type of coupler is commonly used with almost any open guide mounted on a flat surface.

The coupler is usually narrow band because the relative phase of the even and odd modes is highly frequency dependent. The different phase velocities of the two modes lowers directivity performance.

The results of theoretical and experimental work done with these guides are shown in Fig. 5.5 [16,22]. The performance of these couplers is good for many guides and agrees closely with theoretical predictions.

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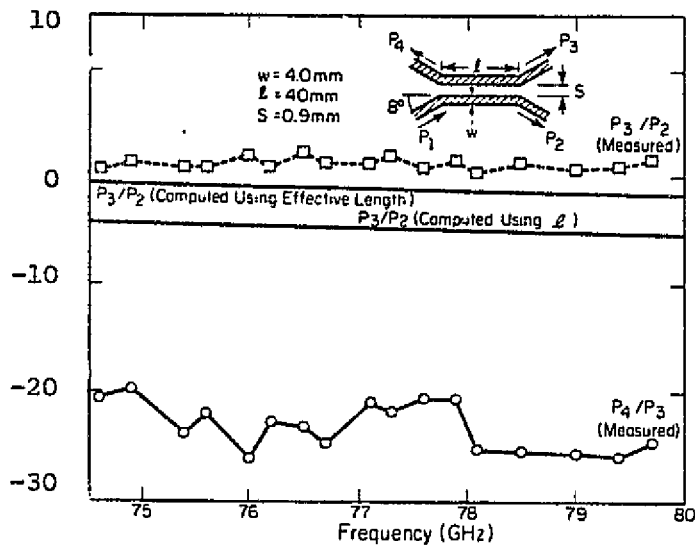


Figure 5.5 Performance of Experimental Distributed Coupler Constructed from Inverted Strip Guide. (Power Ratios vs. Frequency)

(Reproduced from Rudokas, R., and Itoh, T. "Passive Millimeter-Wave IC Components Made of Inverted Strip Dielectric Waveguides." IEEE Trans. MTT, Vol. MTT-24, No. 10, (Oct. 1976), pp. 978-81).

### Aperture Coupling

Aperture coupling is commonly used with closed wall guides, and for these a number of configurations have been used successfully. In these couplers two guide sections share a common wall and fields in the main guide excite fields in the second guide through a number of apertures in this wall. Coupling can be controlled by the size, shape, number and placement of the apertures or by the angle between two guides in the plane of the common wall.

For millimeter use such couplers have been made using WR-15 rectangular guides [28]. Electroforming techniques were used to keep aperture wall thickness small and standard techniques were used in the design. The performance in the millimeter range was as good as at the lower frequencies.

Similar techniques have been used to make couplers for dielectric image guides [29] and this technique could be extended to other millimeter guides that are constructed on a ground plane such as shielded image and inverted strip guides. It may not always be desirable to have a portion of the circuit on both sides of the ground plane as required by these couplers and since these guides are usually designed to keep ohmic loss at a minimum by keeping strong fields away from the ground plane, strong coupling is not always possible.

At frequencies where the coupling wall thickness is approximately  $\lambda_g/4$ , multiple branch couplers can be constructed by

machining the branch lines as part of the wall, avoiding the electroforming of thin coupling walls [28]. For example, in the 88-96 GHz range WR-10 guide has a thickness of .040 inches which is about  $\lambda_g/4$ .

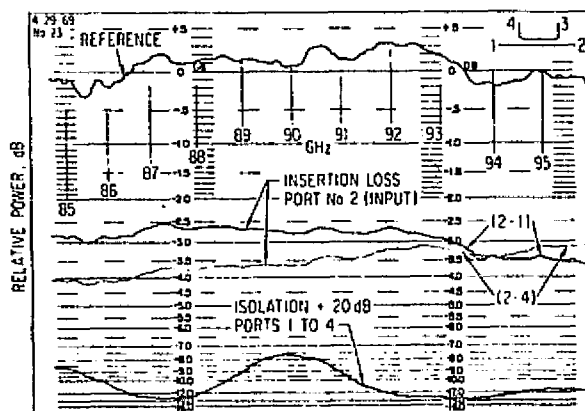
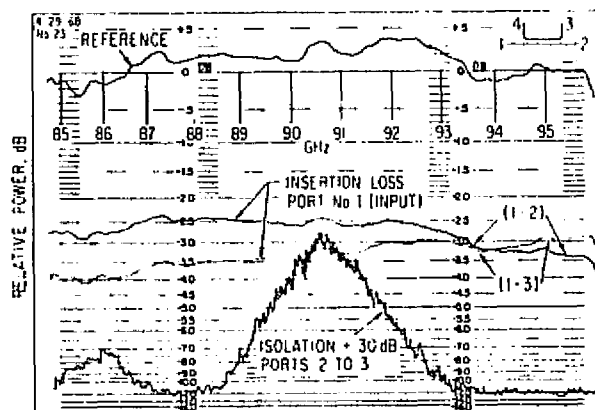
The results of experimental designs for directional couplers used at millimeter wave frequencies are shown in Fig. 5.6 [28,29]. The results for rectangular aperture and multiple branch couplers are as good as those achieved throughout the microwave band. Indeed, it has been shown experimentally that many of the components in rectangular guides used at microwave frequencies are just as effective when scaled directly to the millimeter-wave range down to wavelengths of 1 mm [15].

#### 5.4 Ferrite Based Devices

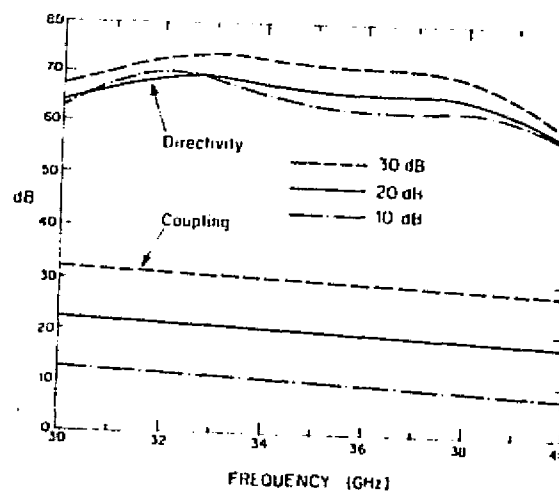
Many useful devices are based on non-reciprocal gyromagnetic effects when microwave fields interact with ferrites. These devices are grouped into two classes: those operating in the region of ferromagnetic resonance, such as isolators, harmonic generators, filters, switches and resonators, and those operating well away from resonance such as Faraday rotation isolators, phase-shifters, circulators, phase-shift-type switches and filters.

In millimeter-wave applications there are several problems which make design more difficult than a direct scaling approach. One problem is the difficulty in the fabrication of very small





(a)



(b)

Figure 5.6 Experimental Performance of Directional Couplers.

a) Multiple Branch 3dB Coupler in Rectangular Guide.

b) Aperture Coupler in Dielectric Image Guide

(Figure (a). Reproduced from Bottjer, Martin F., and King Howard. "Top-Wall and Branch-Waveguide Hybrids for Millimeter Wavelengths." IEEE Trans. MTT, Vol. MTT-20, No. 2 (Feb. 1972), pp. 182-84. Figure (b) Reproduced from Bahl, Inder J., and Bhatia, Prakash. "Aperture Coupling Between Dielectric Image Lines." IEEE Trans. MTT, Vol. MTT-29, No. 9 (Sept. 1981), pp. 891-96).

ferrite parts which is especially costly.

Secondly, the ferromagnetic resonance absorption peak for isotropic, spherically shaped samples occurs at an angular frequency

$$\omega = \gamma H_0 \quad (5.5)$$

where  $\gamma$  is the gyromagnetic ratio (usually 2.8 MHz/oersted) and  $H_0$  is the applied dc magnetic field. At millimeter-wave frequencies the applied fields are inconveniently large, 10,000 oersted at 28 GHz. Also the difference between magnetic susceptibilities ( $X_+ - X_-$ ) decreases as frequency increases and thus limits the performance of Faraday rotation and phase shift devices.

New materials have been developed which are highly anisotropic and have an internal anisotropy field,  $H_a$ , which is large enough to require little or no externally applied magnetic fields since

$$\omega = \gamma(H_0 + H_a)$$

The materials are the hexagonal ferrites and they have been used at frequencies up to 55 GHz with applied magnetic fields of only 4000 oersteds [30].

As described in the chapters on the various guides, certain guides have a propagation mode with a region of elliptical magnetic field polarization. These guides, which include the closed metallic guides and the slotline and coplanar guides are suited for

use with these ferromagnetic materials to produce the various devices listed.

### 5.5 System Designs

As noted earlier, many millimeter-wave circuits use hybrid techniques. A typical receiver might use a dielectric rod antenna, a dielectric guide coupler to introduce local oscillator power and then use a transducer to the  $TE_{10}$  mode of a rectangular guide and to the solid state detector. The advantages of each guide type is utilized to give good performance at low cost. Usually, though, smaller sizes and better operation is achieved by using a single design technique and avoiding the guide transducers when the guide type is changed in a single circuit.

This single guide design technique is also the method desired in monolithic millimeter-wave integrated circuits. As fabrication techniques permit the miniaturization required, monolithic circuits will bring to millimeter wave frequencies the advantages realized at microwave frequencies, namely, low cost batch production, improved reliability and reproducibility and small size and weight.

## CHAPTER 6

### CONCLUSIONS

This paper has discussed transmission systems for use in the millimeter wave frequency range.

The application of existing microwave structures such as the standard rectangular and circular closed metallic guides to transmission of millimeter wave frequencies will remain attractive due to the complete field confinement achieved by these guides and the ease with which existing design techniques can be scaled to the higher frequencies. However, the skin effect increases the losses in these guides with increased frequency, and the reduced cross section requires strict tolerances to insure single-mode propagation and eliminate surface roughness which can further increase losses. The special manufacturing techniques such as precision machining, electroforming and surface polishing which are used to meet these strict tolerances are labor-intensive so that guides and the devices formed from these guides are expensive. Systems based on closed metallic guide structures cannot be mass produced easily by batch process methods and overall system size is usually large since the guides require an external support structure. Therefore, these guides may be best suited for carrying millimeter wave signals over some

distance and the anomalous, low-loss  $TE_{01}$  mode in circular waveguide has the best potential for millimeter wave trunk communication lines.

Planar guides which have reduced size, weight and cost over closed metallic guides at microwave frequencies have the same potential advantages when their use is extended into the millimeter frequency range. The method of applying the metallization layer to a substrate to form the guiding structure, such as etching, makes large scale batch processing possible. This decreases the cost of systems such as those used for millimeter wave imaging or radar where many similar circuits are required.

Planar devices can be more easily used with solid state devices than closed metallic guide, since physical size of these guides facilitates device mounting, and in most cases isolated ohmic contacts required for the d.c. biasing of devices are available. Again, reduced size and ohmic loss in the millimeter wave frequency range become larger and increase the manufacturing difficulties and cost of these guides compared to the similar structures which are used at the lower microwave frequencies.

Dielectric guides, which operate on the basis of optic-like reflection at a series of dielectric interfaces to confine transmitted energy, are uniquely suited to millimeter wave application. The reduced amount of conducting material in the guide decreases ohmic losses and the quasi-optical method of

operation increases cross-sectional dimensions over those of closed metallic and planar guides at a region in the spectrum where increased size is an advantage. Tolerances in the construction of these guides are not as strict as the guides previously described and ease of manufacture and reduced cost can be expected from the use of these guides. Since these guides can be formed on planar substrates, they can be used for millimeter wave integrated circuits, but the lack of isolated ohmic contacts makes the biasing of solid state devices problematic. The fields of these guides can extend far into the surrounding medium and this increases the radiation loss at discontinuities and can lead to cross talk between adjacent circuits on the same plane. Other guides suffer from leakage due to mode conversion which can lead to these same problems. Guide design can reduce some of these negative effects and as the techniques for the analysis of dielectric guides become as well known as the closed metallic and planar guides, further circuit applications in the millimeter wave frequency region will be found. Presently, these guides are used in hybrid structures which utilize their unique properties.

All of these guides are suited to use in the construction of passive components and design and configuration of distributed circuit elements are similar for each guide. The effectiveness of a particular guide for application in construction of milli-

meter wave devices is related to the electrical performance; such as, losses, dispersion and impedance characteristics of that guide.

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The literature on microwave and millimeter wave transmission systems and components is quite extensive and extends back over a period of 40 years or more. It was generally felt that a review of all pertinent material within the main body of the report would result in a loss of perspective on the main trends of development that has taken place. It was therefore decided to supplement the main discussion by the simple addition of a bibliography that would direct the reader to various papers that provide more detailed technical information on the characteristics of various transmission systems and components that have been described in the open literature. Even this bibliography is of limited scope but will serve as a useful beginning for anyone wishing to carry out an exhaustive review.

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